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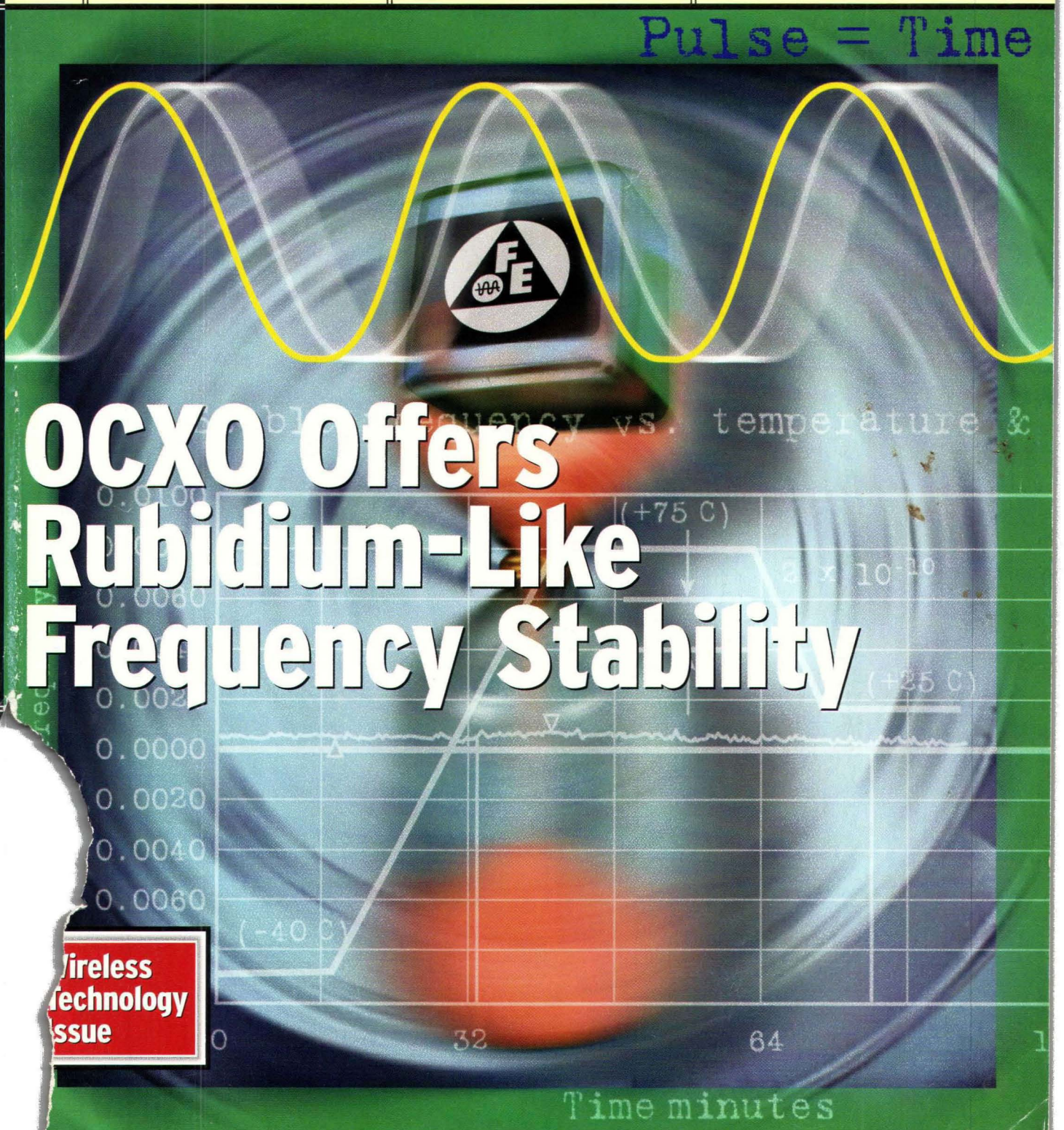
News
Multilayer approach yields SMT circulator

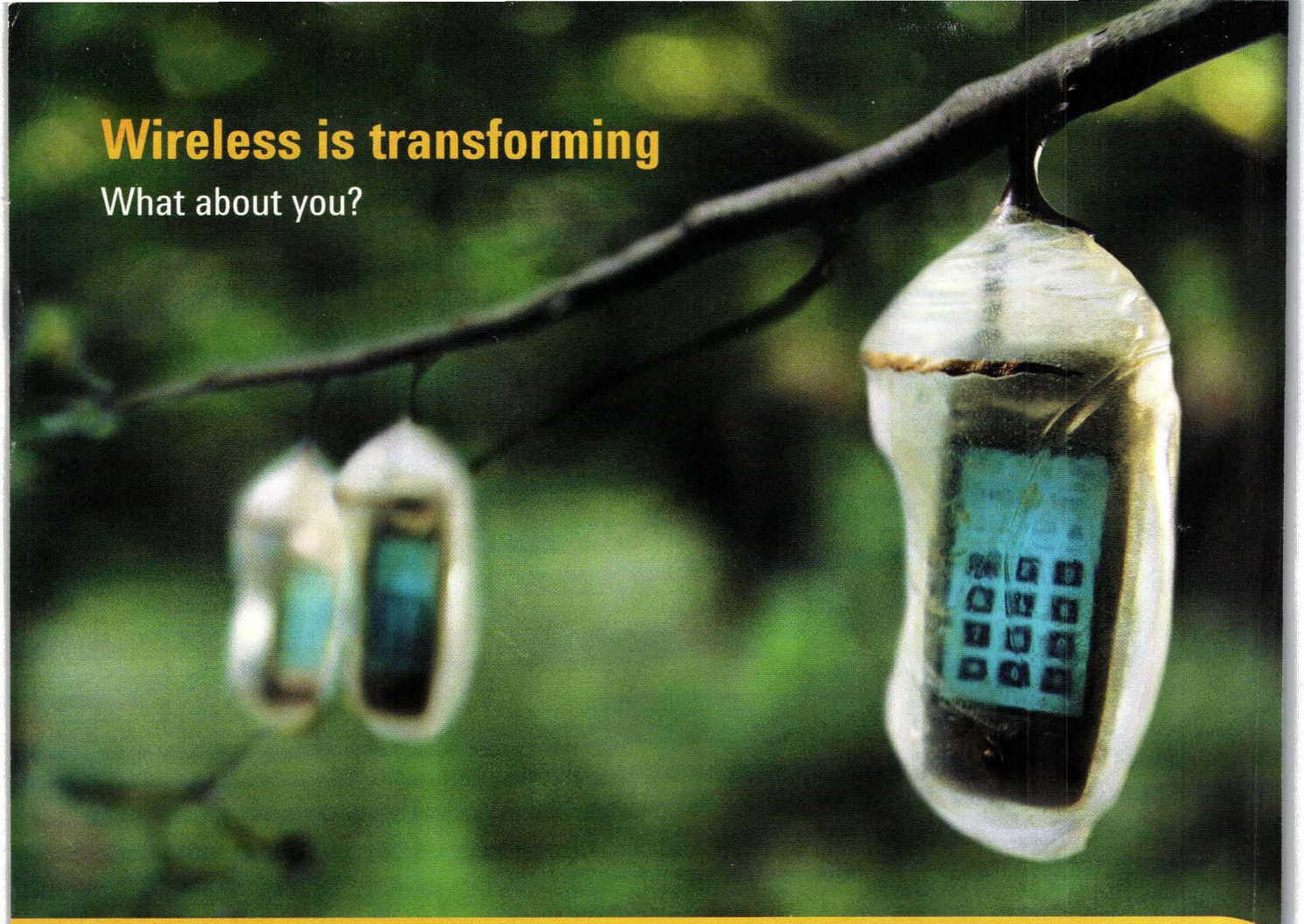
Design Feature
Analyze phase noise in short-range radios

Product Technology
Chipset enables GSM/GPRS/EDGE



OCXO Offers Rubidium-Like Frequency Stability

Wireless Technology Issue



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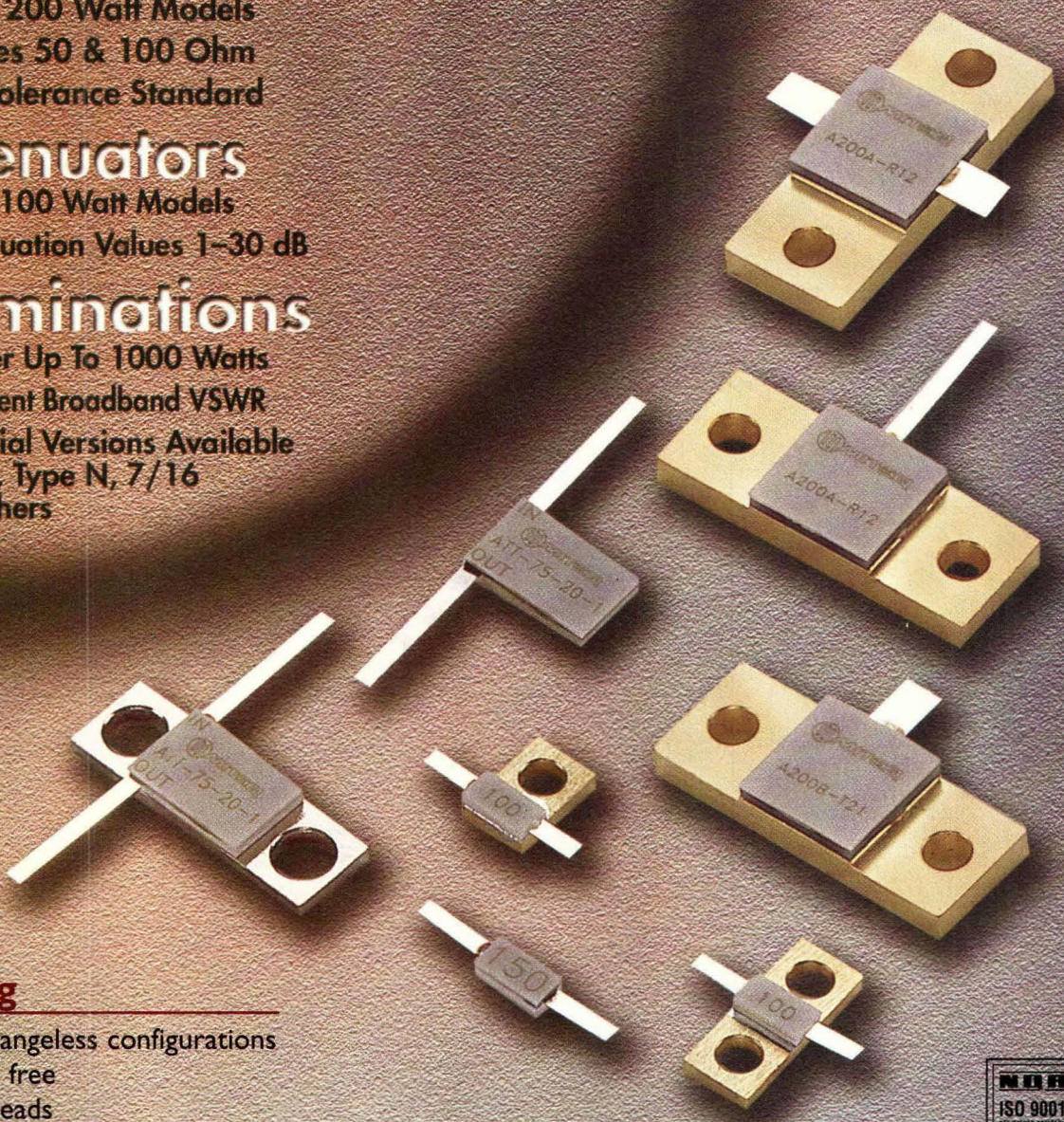

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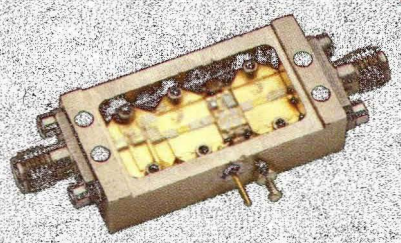

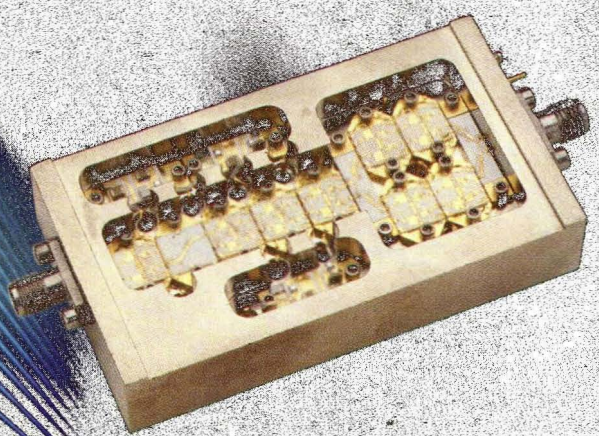

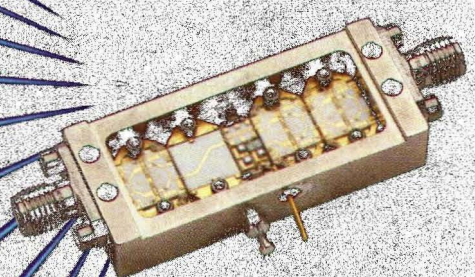

ULTRA BROAD BAND

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA018-203	0.5-18.0	20	5.0	2.5	7	17	2.0:1	250
JCA018-204	0.5-18.0	25	4.0	2.5	10	20	2.0:1	300
JCA218-506	2.0-18.0	35	5.0	2.5	15	25	2.0:1	400
JCA218-507	2.0-18.0	35	5.0	2.5	18	28	2.0:1	450
JCA218-407	2.0-18.0	30	5.0	2.5	21	31	2.0:1	500

MULTI OCTAVE AMPLIFIERS

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA04-403	0.5-4.0	27	5.0	1.5	17	27	2.0:1	550
JCA08-417	0.5-8.0	32	4.5	1.5	17	27	2.0:1	550
JCA28-305	2.0-8.0	22	5.0	1.0	20	30	2.0:1	550
JCA212-603	2.0-12.0	32	5.0	3.0	14	24	2.0:1	550
JCA618-406	6.0-18.0	20	6.0	2.0	25	35	2.0:1	600
JCA618-507	6.0-18.0	25	6.0	2.0	27	37	2.0:1	800

MEDIUM POWER AMPLIFIERS

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-P01	1.35-1.85	35	4.0	1.0	33	41	2.0:1	1000
JCA34-P02	3.1-3.5	40	4.5	1.0	37	45	2.0:1	2200
JCA56-P01	5.9-6.4	30	5.0	1.0	34	42	2.0:1	1200
JCA812-P03	8.0-12.0	40	5.0	1.5	33	40	2.0:1	1700
JCA1218-P02	12.0-18.0	22	4.0	2.0	25	35	2.0:1	700

LOW NOISE OCTAVE BAND LNA'S

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-3001	1.0-2.0	40	0.8	1.0	10	20	2.0:1	200
JCA24-3001	2.0-4.0	32	1.2	1.0	10	20	2.0:1	200
JCA48-3001	4.0-8.0	40	1.8	1.0	10	20	2.0:1	200
JCA812-3001	8.0-12.0	32	1.8	1.0	10	20	2.0:1	200
JCA1218-600	12.0-18.0	45	2.0	1.0	10	20	2.0:1	250

NARROW BAND LNA'S

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-1000	1.2-1.6	25	0.75	0.5	10	20	2.0:1	80
JCA23-302	2.2-2.3	30	0.8	0.5	10	20	2.0:1	80
JCA34-301	3.7-4.2	30	1.0	0.5	10	20	2.0:1	90
JCA56-401	5.4-5.9	40	1.0	0.5	10	20	2.0:1	120
JCA78-300	7.25-7.75	27	1.2	0.5	13	23	2.0:1	120
JCA910-3000	9.0-9.5	25	1.2	0.5	13	23	1.5:1	150
JCA910-3001	9.5-10.0	35	1.2	0.5	13	23	1.5:1	150
JCA1112-3000	11.7-12.2	27	1.1	0.5	13	23	1.5:1	150
JCA1213-3001	12.2-12.7	25	1.1	0.5	10	20	2.0:1	200
JCA1415-3001	14.4-15.4	35	1.4	1.0	14	24	2.0:1	200
JCA1518-3001	18.1-18.6	25	1.2	0.5	10	20	2.0:1	200
JCA2021-3001	20.2-21.2	25	2.0	0.5	10	20	2.0:1	200

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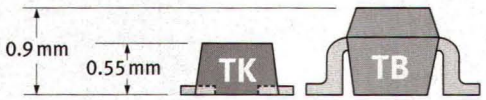


UPG2008TK *SOT-363*

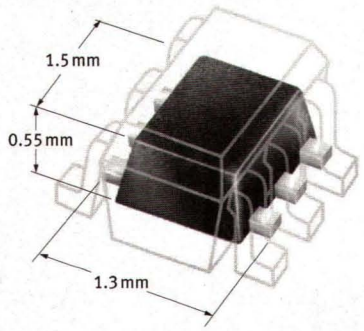
Meet the UPG2008TK. It's footprint is less than half that of a standard SOT-363 switch. Plus its leads are flat and recessed into the base of its package, giving it that super low profile your miniaturized portable products demand. Optimized for transmit/receive switching, antenna diversity switching, and power level control, the UPG2008TK is the latest in our growing family of GaAs MMIC SPDT switches. See them all on our web site at:

www.cel.com/spdt.asp

GaAs MMIC SPDT Switches

Part Number	Insertion Loss @ 1.0 GHz	P_{IN} Power Handling	Control Voltage	Package	100K Price	Description
UPG2008TK	0.4 dB	+25 dBm @ 1.0 dB	2.8 V	TK	45¢	*World's Smallest*
UPG2009TB	0.25 dB	+34 dBm @ 0.1 dB	2.8 V	TB	78¢	*High Power, No Compromises*
UPG2006TB	0.35 dB	+20 dBm @ 1.0 dB	1.8 V	TB	54¢	*Low Voltage, Great Specs*
UPG158TB	0.3 dB	+25 dBm @ 0.1 dB	3 V	TB	39¢	*Good Specs, Great Price*
UPG152TA	0.4 dB	+30 dBm @ 1.0 dB	3 V	TA	29¢	*Low Cost 3V Switch*

Applications for CEL's NEC switches include: Bluetooth, PCMCIA Cards, Cordless Phones, Digital Cellular, PCS, WLAN, and WLL.



NEC's new TK package has a 50% smaller footprint than the TB (SOT-363) package switches.

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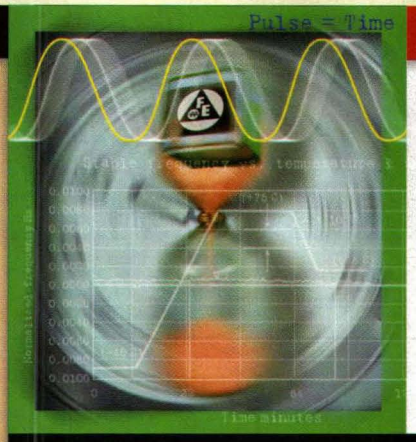

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Microwave Radios	9.0 - 15	-8	17/9.0	+13	8	50	HMC412MS8G
PCS & ISM	1.3 - 2.5	+8	21/29	+10	10	30	HMC421QS16
MMDS & WLL	1.2 - 2.5	-8	15/23	0	8	30	HMC422MS8
Cellular ISM	0.6 - 1.3	-8	15/23	0	8	35	HMC423MS8

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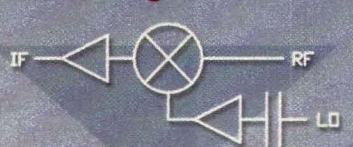

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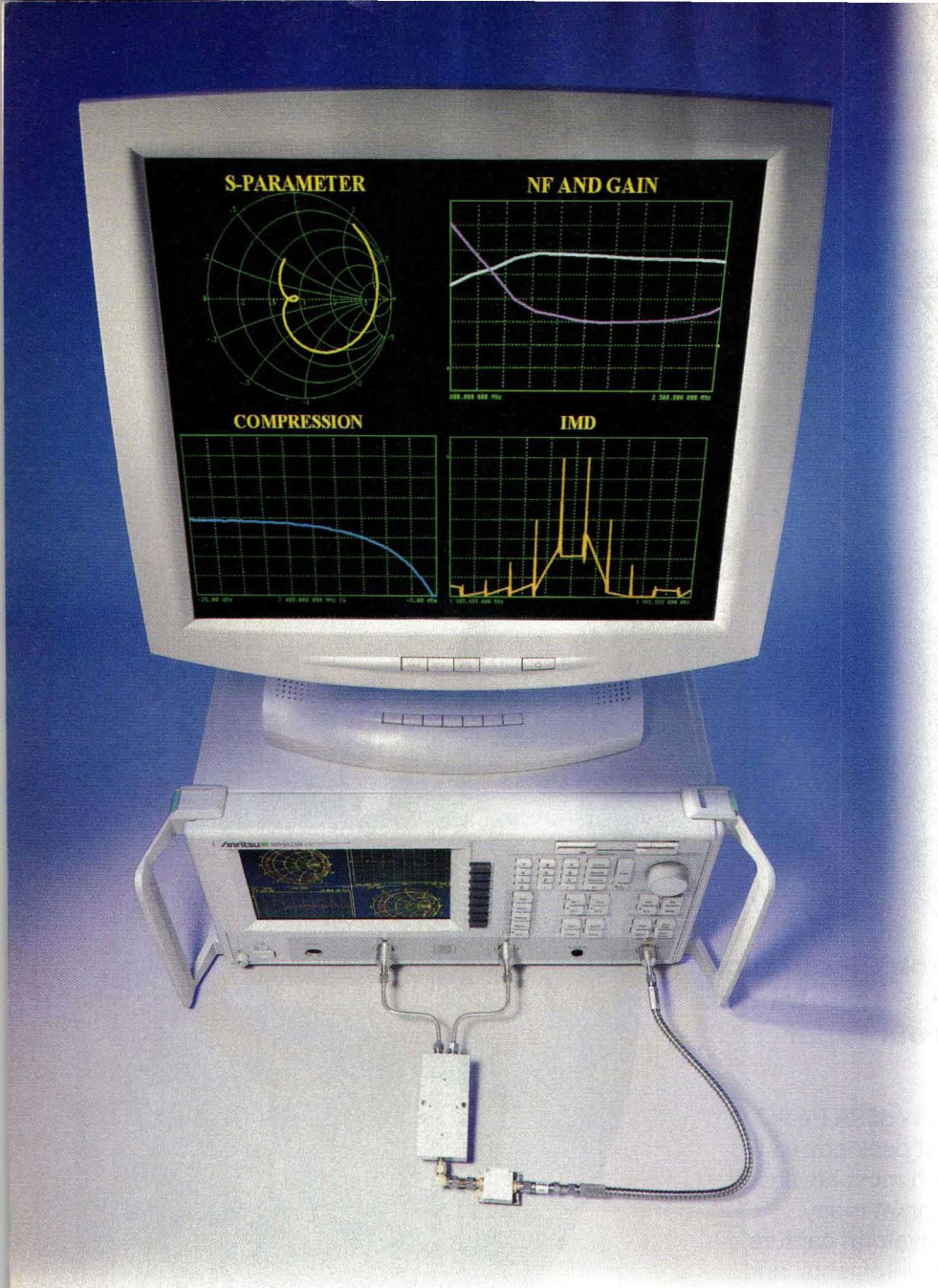

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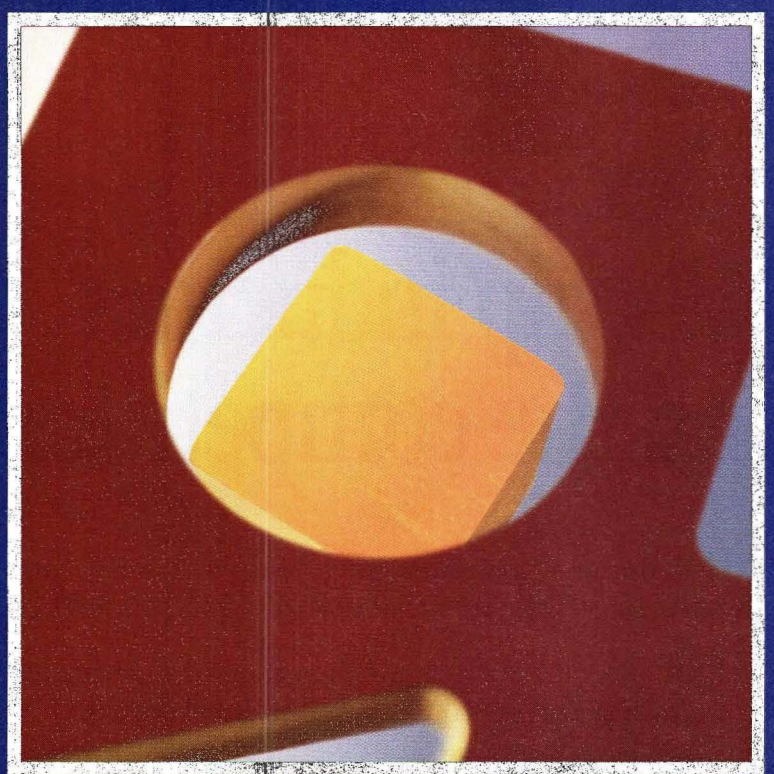

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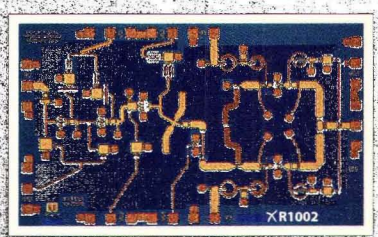

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Incorrect Statements

(continued from last month)

▶▶THE REFLECTED WAVE is not "absorbed" by the source. What happens is that the reflected wave is re-reflected at the source while adding to the incident (forward) power on the transmission line. The power absorbed by the load is equal to the difference between the incident power and the reflected power. This effect can be readily measured by inserting a directional wattmeter in the transmission line at any point. The power delivered by the source into the transmission line will equal the power absorbed by the load, while the incident power will be greater than the power absorbed at the load by an amount equal to the reflected power.

Suppose a matching network is inserted in the transmission line, and this network is adjusted to provide a conjugate match between the two sections of transmission line. Toward the source and looking into the transmission line, the impedance will be Z_o (assuming that Z_s and Z_o are equal). Toward the load, the impedance will be Z_o, Z_L, and the length of the transmission line. The output impedance of the matching network will be the complex conjugate of this.

If the wattmeter is inserted in the transmission line on the source side of the matching network, the incident power will equal the available power of the source, and there will not be any reflected power. If the wattmeter is moved to the load side of the network, the incident power can be seen to increase by an equal amount to the reflected power now on the like. The difference will equal the incident power measured on the source side of the network. All of the reflection loss that occurs at the load is compensated for by an equal reflection gain at the matching network. The incident power minus the reflected power will exactly equal the power delivered from the source. This is according to the conjugate matching theorem which states that under conjugate-matched conditions, all available source power will be absorbed by the load, and that if a conjugate match exists at one location in a network, then a conjugate match exists at all locations in the network.

Walter Maxwell then says, "For the case where Z_s is not equal to Z_o, a portion of the reflected wave from the load is re-reflected from the source back toward the load." See the previous comment regarding quarter-wave matching sections.

Maxwell explains all of this very clearly in his book, *Reflections II*. This is highly recommended reading.

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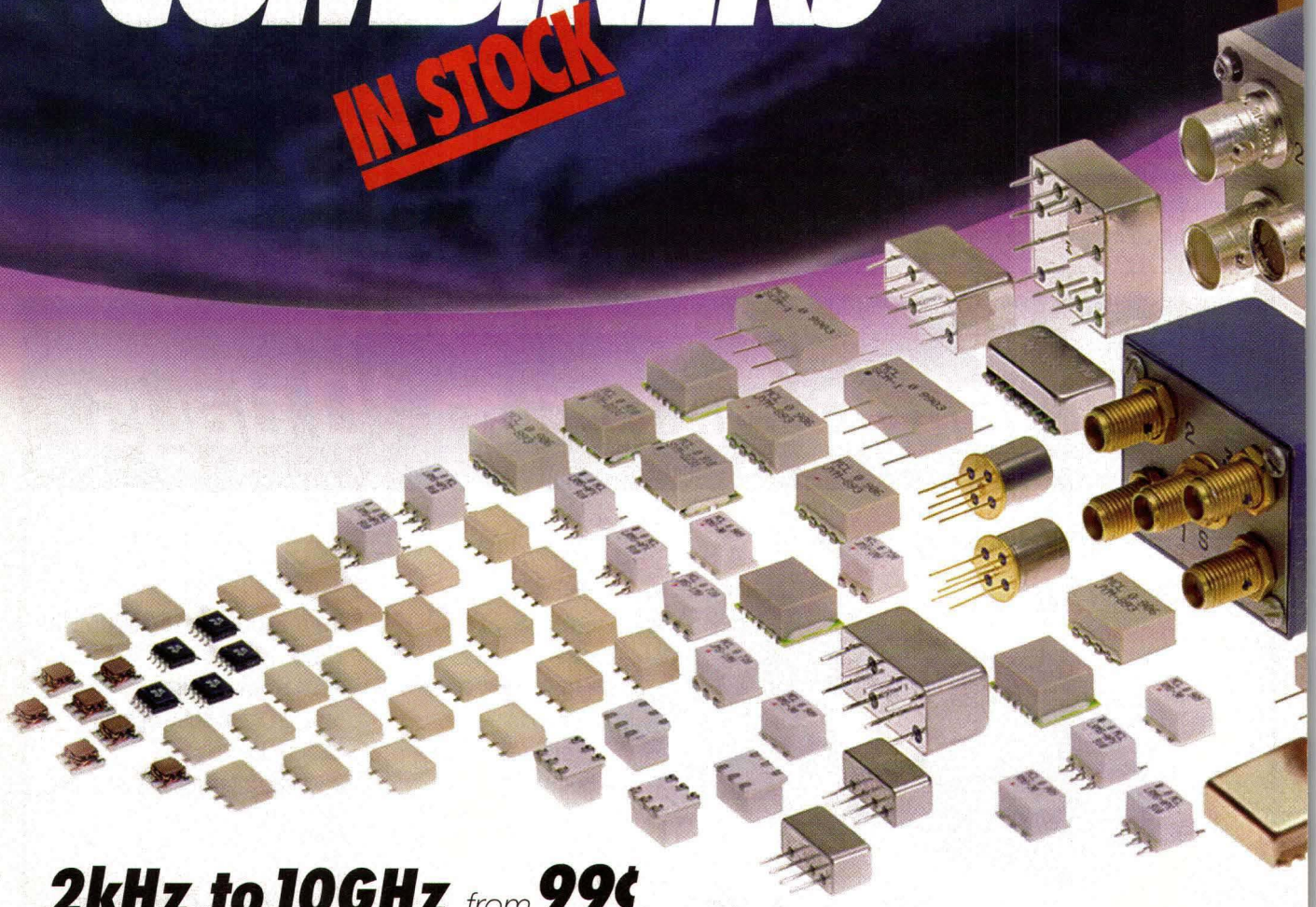

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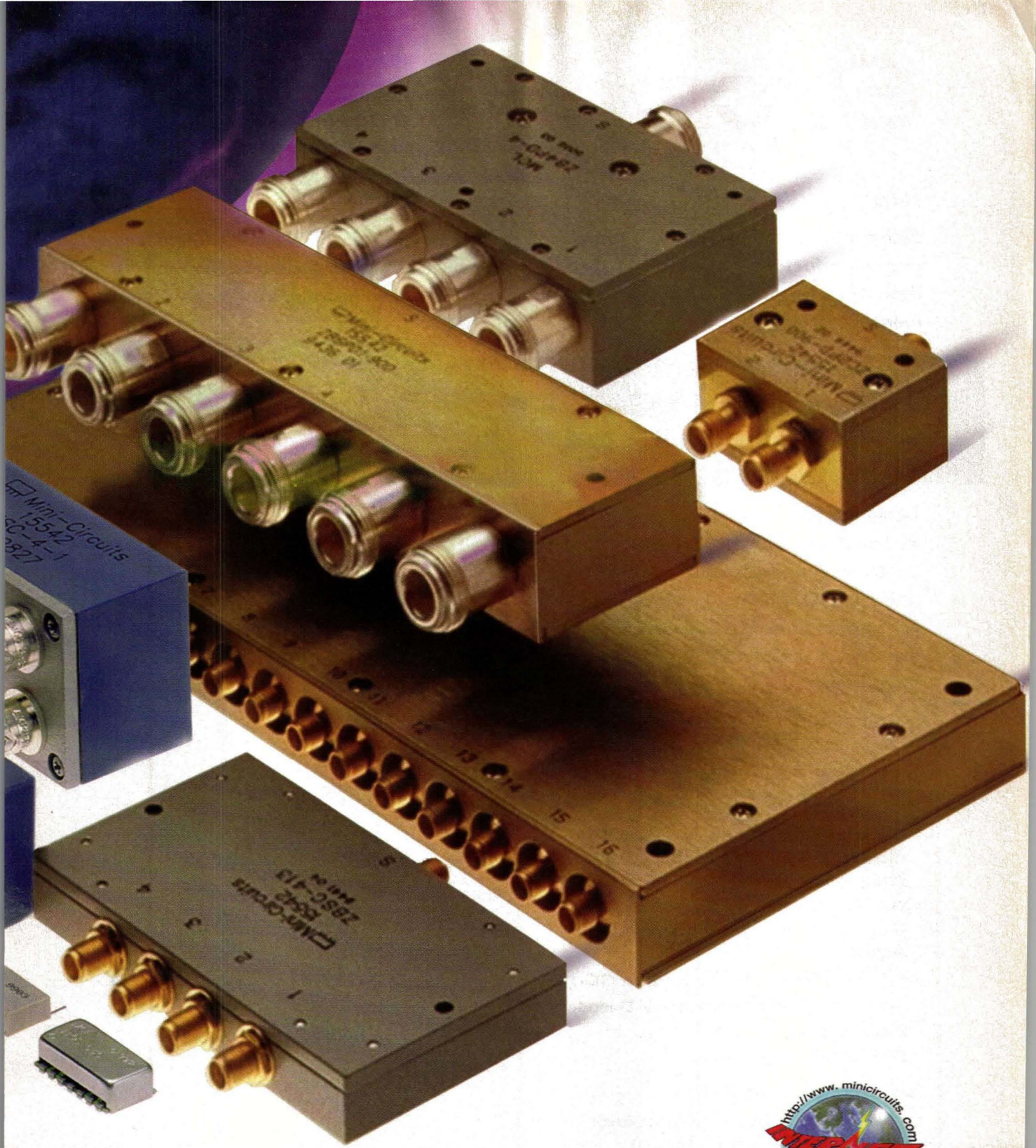

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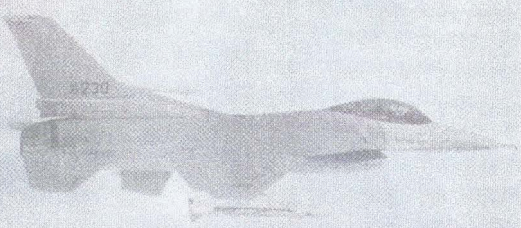

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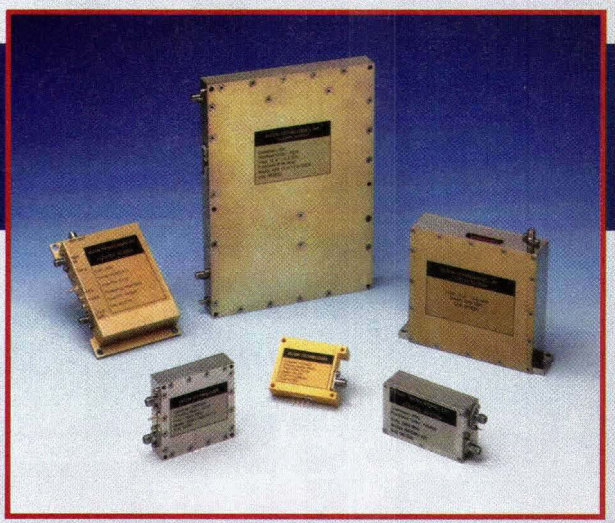

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Helping Advance Military Designs

Military technology is once again of interest to high-frequency designers and manufacturers. To those who fear terrorism, the presence of military peacekeeping forces provides a source of comfort. To microwave designers, the military arena has once again become a hotbed of opportunities and challenges.

The events of September 11, 2001 have taught us that our intelligence capabilities are far behind the technology curve. While each successive generation of cellular telephone strives for more features and better quality, the methods employed to accomplish this—such as spread spectrum and CDMA techniques—make it more difficult to maintain effective surveillance. The decade of the 1990s, which saw most of the industry target commercial communications markets, resulted in a refinement of many former military technologies for commercial use, but without corresponding improvements in surveillance.

The events of September 11, 2001 have taught us that our intelligence capabilities are far behind the technology curve.

If our military forces are to maintain their edge against advancing technologies, they must stay one step ahead. Is there a simple solution for this, other than having key technology companies contributing design time and resources to such organizations as the NSA and the FBI? Perhaps not. But there is a way to steer the technological tide back in favor of the military, and that is by boosting awareness, having a willingness to work with military customers, and by listening to the needs of the military customer.

One small way to increase this awareness is by a sharing of ideas on military design: the basic goal of the Military Electronics Show (MES). The second MES (the inaugural event was held last April in Baltimore) is sponsored by *Microwaves & RF* magazine. It is scheduled for September 24-25, 2002 at the Baltimore Convention Center (Baltimore, MD). This editor will serve as the Technical Program Chairman for the event, with the humble goal of trying to recruit speakers and technical moderators willing to discuss issues impacting the design of components, systems, and measurement techniques for military applications, such as avionics systems, direct-finding Rxs, ECM systems, EW systems, fiber-optic systems, jammers, RPVs, SATCOM systems, secure communications systems, and surveillance Rxs.

Terrorism can be stopped, with the help of a strong military. If you would like to be part of helping the military advance, then plan to join us in Baltimore this September, as a speaker or an exhibitor. Either way, we will try to make our peacekeeping forces stronger and wiser.

Jack Browne

Publisher/Editor

Microwaves & RF

A Penton Publication

HIGH-SPEED ELECTRONICS GROUP
Group Publisher Craig Roth, (201) 393-6225 • *croth@penton.com*
Publisher/Editor Jack Browne, (201) 393-6293 • *jbrowne@penton.com*
Managing Editor Peter Stavenick, (201) 393-6028 • *pstavenick@penton.com*
Senior Editor Gene Heftman, (201) 393-6251 • *gheftman@penton.com*
Associate Managing Editor John Curley, (201) 393-6250 • *jcurley@penton.com*
Special Projects Editor Alan ("Pete") Conrad
Copy Editor Mitchell Gang • *mgang@penton.com*
Editorial Assistant Dawn Prior • *dprior@penton.com*
Contributing Editors Andrew Laundrie, Allen Podell

MANUFACTURING GROUP
Director Of Manufacturing Ilene Weiner
Group Production Director Mike McCabe
Customer Service Representative
Dorothy Sowa, (201) 393-6083, FAX: (201) 393-0410
Production Coordinator Judy Osborn, (201) 393-6258

ART DEPARTMENT
Art Director Armand Veneziano • *aveneziano@penton.com*
Group Design Manager Anthony Vitolo • *tvitolo@penton.com*

CIRCULATION CUSTOMER SERVICE (LIVE) (847) 647-6657
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EDITORIAL OFFICE
Penton Media, Inc.
611 Route #46 West, Hasbrouck Heights, NJ 07604
Phone: (201) 393-6286, FAX: (201) 393-6227

PENTON TECHNOLOGY MEDIA
President David B. Nussbaum
Vice President, Finance Keith DeAngelis
Director, Information Technology Steven Miles
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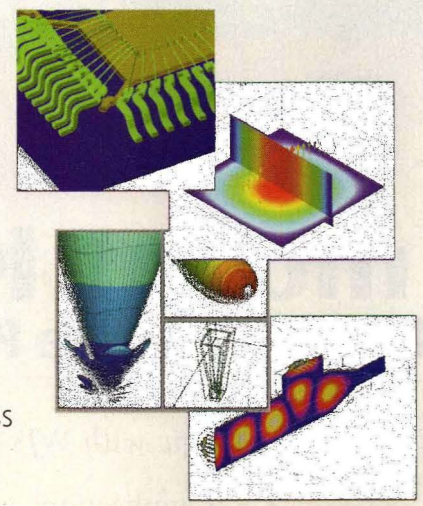

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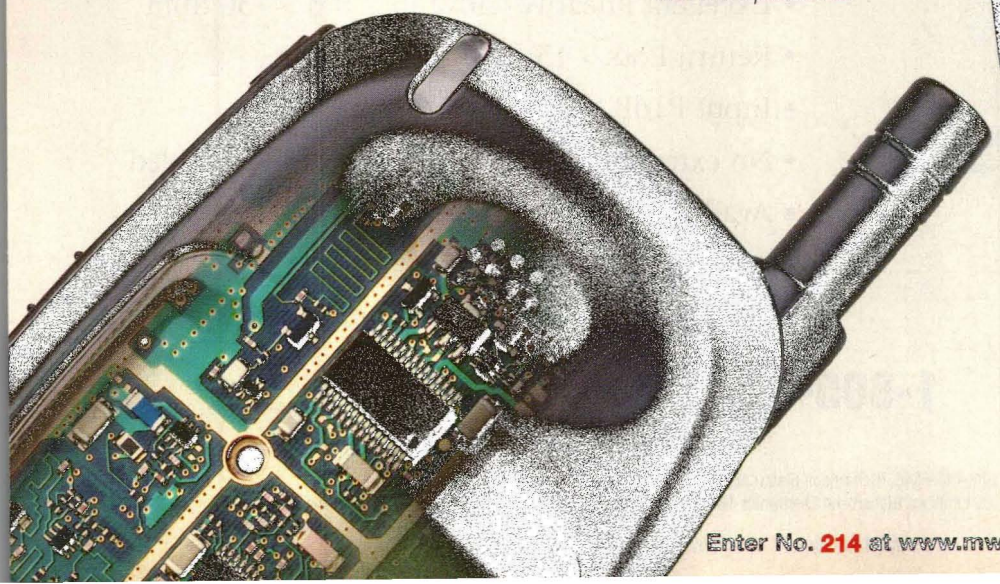

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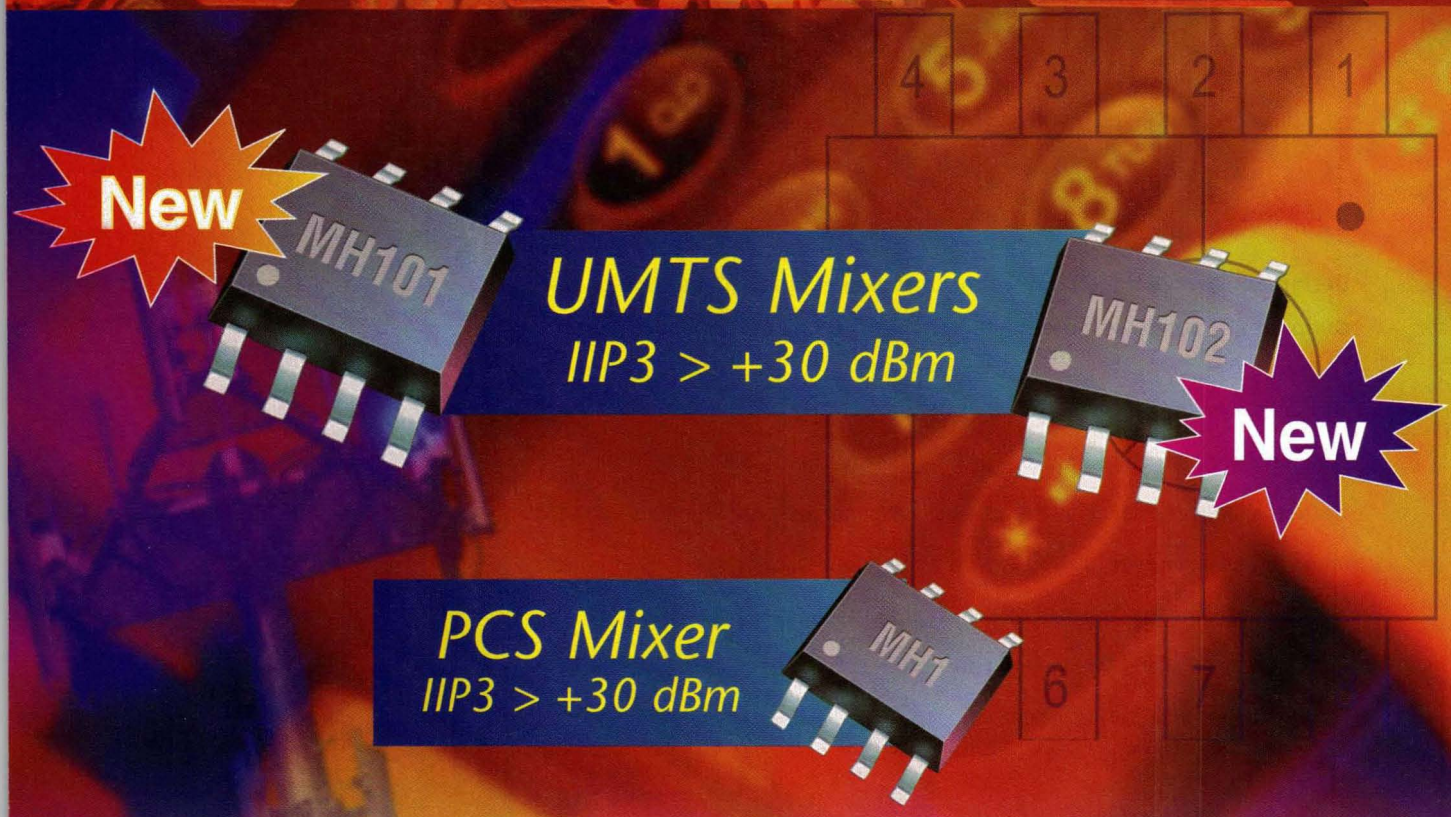

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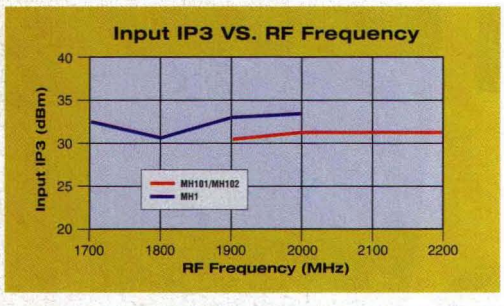

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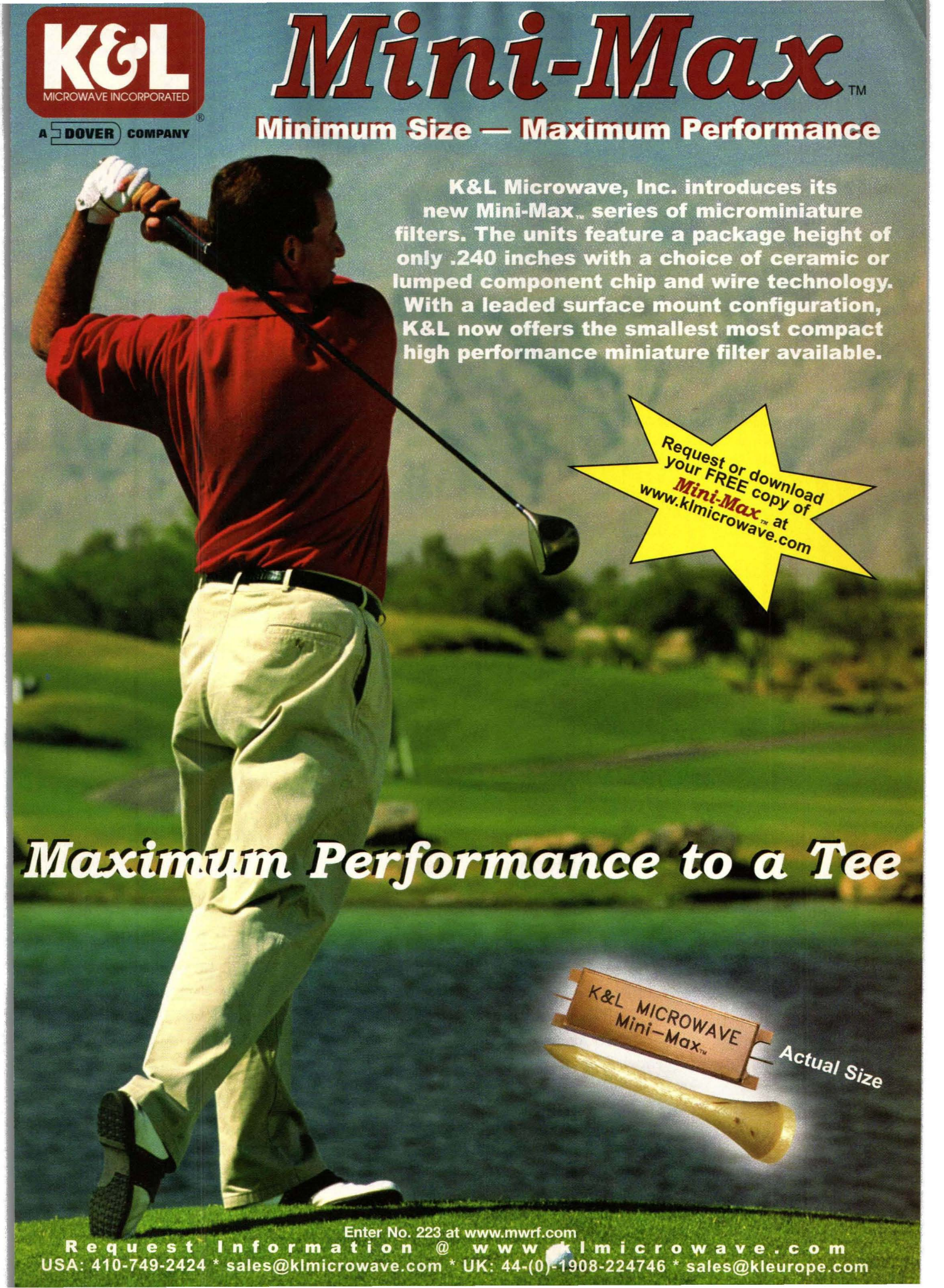
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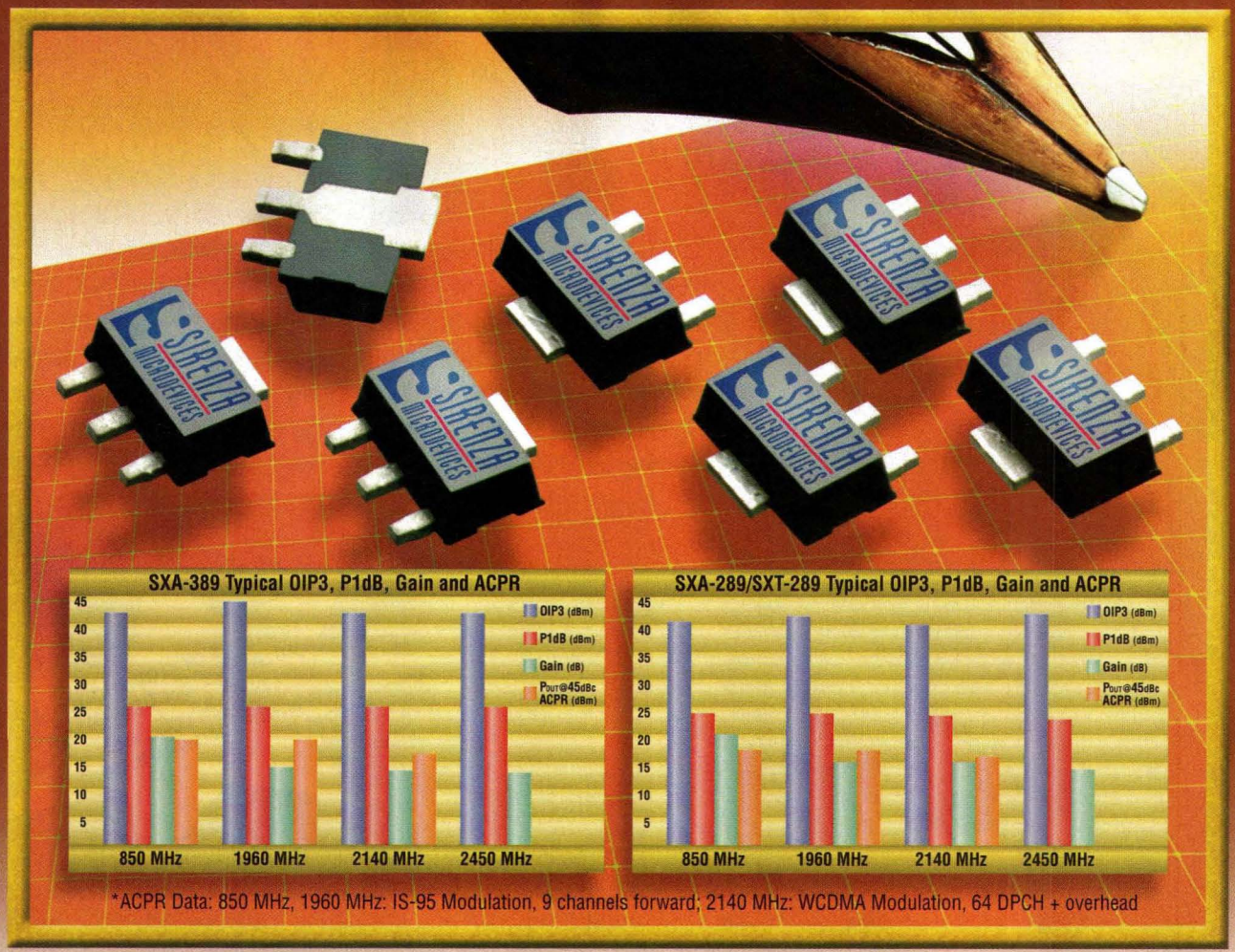

*ACPR Data: 850 MHz, 1960 MHz: IS-95 Modulation, 9 channels forward; 2140 MHz: WCDMA Modulation, 64 DPCH + overhead

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the front end

News items from the communications arena.

Electronic Polymer Products Market Will Expand To $3.5 Billion In 2005

CLEVELAND, OH—The demand for electronic polymer products is forecast to expand 6.4 percent per year to $3.5 billion in 2005 **(see table)**. This will require 485 million pounds of resin valued at $1.6 billion. Growth in electronic polymer products has lagged gains in electronic components over the last decade due to intense pricing pressure and the tendency of chip manufacturers to outsource their chip packaging and assembly activities to lower-cost companies in Asia. US firms have responded to this ongoing threat by investing heavily in research-and-development (R&D) programs regarding new packaging technologies and higher-value processes and products. The shift to new packaging technologies, particularly direct chip attachment, is intensifying the need for close collaboration between chip makers and back-end operations, which should favor domestic firms over offshore concerns. These and other trends are presented in *Electronic Polymers*, a study from The Freedonia Group, Inc., a Cleveland-based industrial market-research firm.

Electronic Polymer Products Demand (Millions of dollars)

ITEM	1995	2000	2005	PERCENTAGE OF ANNUAL GROWTH 00/95	05/00
Electronic polymer products demand	1820	2532	3450	6.8	6.4
Laminates	987	1386	1795	7.0	5.3
Packaging materials	604	802	1070	5.8	5.9
Adhesives and underfill	139	210	320	8.6	8.8
Other products	90	134	265	8.3	14.6
Electronic polymer demand	791	1129	1630	7.4	7.6

Source: The Freedonia Group, Inc.

The industry's most rapid annual gains are forecast for low-k dielectric polymers, with demand reaching $75 million in 2005 from negligible levels in 2000. Low-k dielectric polymers (or polymers exhibiting a low dielectric constant) are challenging traditional silicon dioxide in crucial interlayer-dielectric (ILD) applications in semiconductor fabrication.

Wireless 3G Network Is Showcased In Public Demonstration

LAS VEGAS, NV—Sprint demonstrated its first live, public high-speed wireless connection to the Internet via the Sprint Third Generation Network and Third Generation Sprint PCS Wireless Connection Card℠ recently at the Consumer Electronics Show in Las Vegas. The demonstration was part of the Sprint Third Generation Experience, an educational mobile showcase that debuted at CES, featuring live and simulated 3G applications, devices, and services that will be available following the nationwide launch of the Sprint Third Generation Network later this year.

The live demonstration featured high-speed wireless access to the Internet, as well as enterprise applications such as e-mail, using a Sprint PCS Wireless Connection Card installed in a laptop and handheld computing device. A digital camera with a PCMCIA slot was also used to send digital images to a hosted website via the Sprint Third Generation Network. The company expects to have 3G Wireless Connection Cards available to customers in time for the nationwide launch of Sprint's Third Generation Network. The 3G card will provide customers with seamless wireless Internet connectivity, connecting notebook or handheld devices, as well as digital cameras, directly to the Sprint Third Generation Network at speeds of up to 144 kb/s.

Sprint has been preparing its network for the migration to 3G for over a year. The company's migration to 3G1X, which consists mainly of simple channel element hardware and software upgrades to base stations, is either complete or under way in other markets in the US.

Smith Amendment Will Aid Tech Sector, Says Secretary Evans

WASHINGTON, DC—Commerce Secretary Donald Evans has called on the US Senate to pass the Smith amendment to the economic stimulus package, according to *Commerce News*.

"We need to do all we can at this time to keep America working and reinvigorate our great economic engine. Accelerating depreciation over the next three years, as the President proposed, could provide meaningful stimulus for our innovators and entrepreneurs. By contrast, the one-year proposal offered by the Majority Leader will do far less for our workers and businesses," Evans says.

"America's technology sector has powered our economy for much of the past decade. We need to help them get back on their feet and to the forefront of our growth once again."

Under the incentive provision, companies that purchase technology assets and place them into service would be permitted to accelerate the expensing of those assets through a one-time depreciation deduction. Many suggest that such a deduction could help reinvigorate the information-technology (IT) industry, which has been struggling since 2000 but remains critical to economic recovery in the US.

"Accelerated depreciation is particularly critical for our technology sector," observes Evans. "America's technology sector has powered our economy for much of the past decade, creating jobs, increasing productivity, and expanding exports. We need to help them get back on their feet and to the forefront of our growth once again. The Smith Amendment will help reignite technology growth."

Secretary Evans additionally expressed hope that the accelerated depreciation proposed by the President, and embodied in the Smith Amendment, will speed infrastructure deployment, especially for high-speed, high-capacity data networks—the broadband Internet.

"We believe the Smith Amendment will also help accelerate the deployment and usage of broadband networks," opines Evans. "Broadband holds great promise for our economy and our society, helping employers revolutionize their processes and business methods. Broadband will also improve education, health care, national security, and entertainment."

Broadband Penetration Is Low In The United Kingdom

LONDON, ENGLAND—Britain's telecommunications regulator has reported a surge in Internet use, with almost half of UK households now online. But Britons are using old technology to get connected, according to a report from Reuters.

In terms of narrowband—using traditional equipment and wires—UK Web connections jumped by 50 percent to a total of 11 million households over the year to November, figures from the Office of Telecommunications (Oftel) show.

Oftel's research also revealed the surprising fact that broadband access—using high-speed cables or equipment—is used by less than one percent of Britain's population. This is one of the lowest penetration rates in the industrialized world.

Oftel had previously expressed concern about the broadband situation, but it has decided that the narrowband system gives some of the best value for money available.

"UK consumers get some of the cheapest, best deals in the world," an Oftel spokesman states.

Oftel says that British Telecommunications plc (BT) was dominant in the wholesale market, but BT was working with other suppliers so that it would ensure that the situation remains competitive.

"We're pleased with the main result of the review," a BT spokesman says. "We're committed to broadening Internet access across the UK."

By 2005, SS7 Network Elements Shipments To Reach $11.5 Billion

NATICK, MA—Recent analysis from Venture Development Corp.'s (VDC) Telecommunications Group reveals that the global market for Signaling System 7 (SS7) network elements leveled off at $9.4 billion in 2000. According to VDC's market study entitled *The Global Market for Signaling Products and Related Equipment in PSTN and Voice over Packet Networks; Volume I: SS7,* shipments of SS7 network elements are expected to increase by just over 4 percent annually to reach approximately $11.5 billion by 2005.

Despite the overall modest growth projections, VDC believes that there are still many opportunities for SS7 vendors, most of which lie in the emerging regions of Asia/Pacific, Latin America, and Rest of World (Middle East and Africa).

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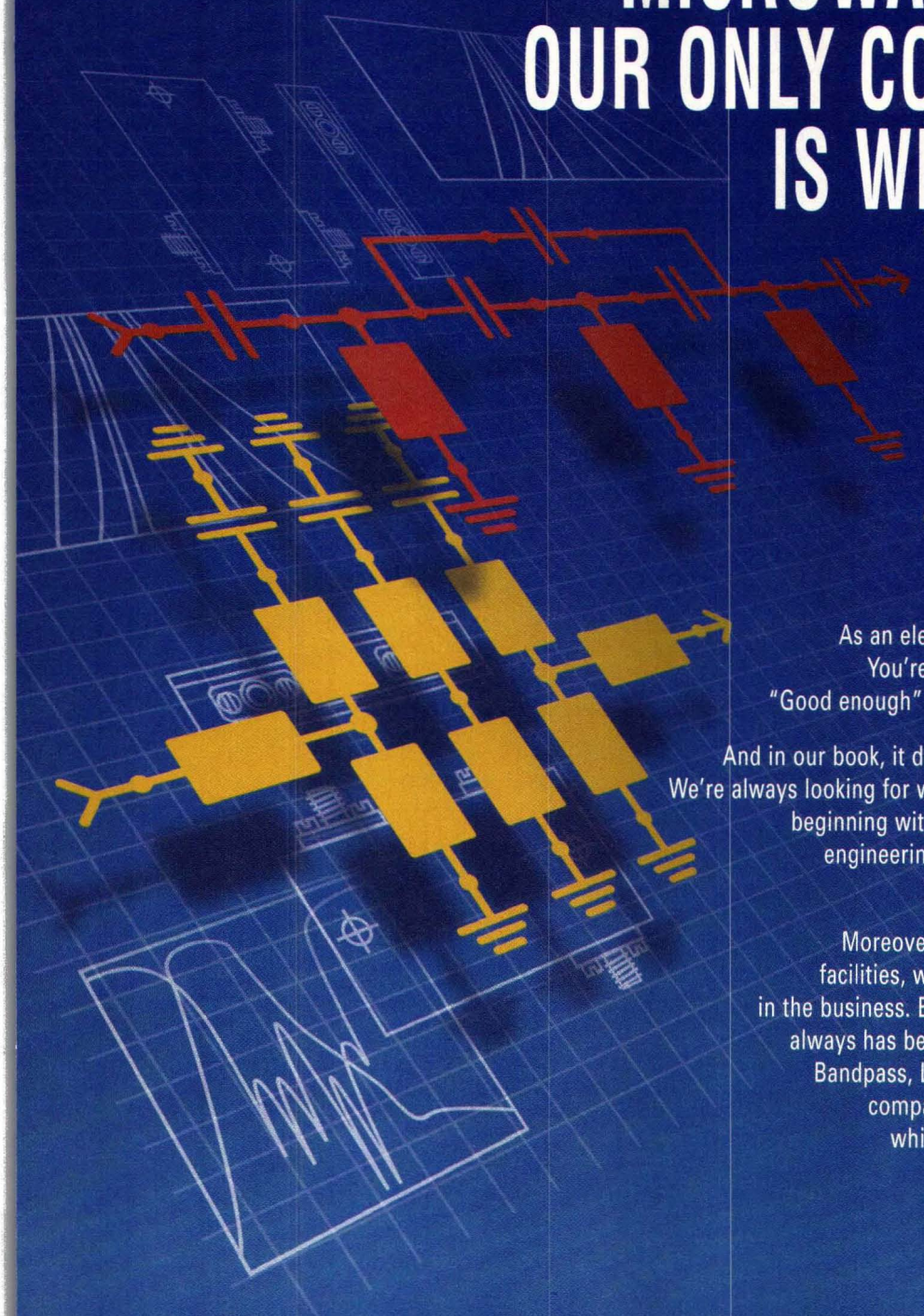

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JS2-00500100-12-5A	0.5 – 1	35	1.2	1	2:1	2:1	5	250
JS2-01000200-045-5A	1 – 2	33	1	0.45	2:1	2:1	5	250
JS2-02000400-045-5A	2 – 4	28	1.2	0.45	2:1	2:1	5	175
JS2-04000800-08-0A	4 – 8	22	1.2	0.8	2:1	2:1	0	150
JS3-04000800-08-5A	4 – 8	30	1	0.8	2:1	2:1	5	175
JS3-04000800-15-5A	4 – 8	30	1	1.5	2:1	2:1	5	175
JS2-08001200-11-5A	8 – 12	15	1	1.1	2:1	2:1	5	150
JS3-08001200-11-5A	8 – 12	25	1	1.1	2:1	2:1	5	175
JS3-08001200-15-5A	8 – 12	25	1	1.5	2:1	2:1	5	175
JS3-12001800-16-5A	12 – 18	23	1	1.6	2:1	2:1	5	175
JS4-12001800-145-5A	12 – 18	30	1	1.45	2:1	2:1	5	200
JS4-12001800-30-5A	12 – 18	30	1	3	2:1	2:1	5	200
JS2-18002600-20-5A	18 – 26	14	2	2	2.5:1	2.5:1	5	100
JS2-18002600-30-5A	18 – 26	14	2	3	2.5:1	2.5:1	5	100
JS3-18002600-20-5A	18 – 26	22	1.8	2	2.5:1	2.5:1	5	175
JS3-18002600-30-5A	18 – 26	22	1.8	3	2.5:1	2.5:1	5	175
JS4-18002600-19-5A	18 – 26	33	1.5	1.9	2:1	2:1	5	200
JS4-18002600-26-5A	18 – 26	33	1.5	2.6	2:1	2:1	5	200
JS2-26004000-35-5A	26 – 40	10	2	3.5	2.5:1	2.5:1	5	100
JS2-26004000-45-5A	26 – 40	10	2	4.5	2.5:1	2.5:1	5	100
JS3-26004000-35-5A	26 – 40	18	2.5	3.5	2.5:1	2.5:1	5	175
JS3-26004000-45-5A	26 – 40	18	2.5	4.5	2.5:1	2.5:1	5	175
JS4-26004000-40-5A	26 – 40	23	2.5	4	2:1	2:1	5	200
JS4-40006000-65-0A	40 – 60	15	3	6.5	2.75:1	2.75:1	0	175
MULTIOCTAVE BAND AMPLIFIERS								
JS2-00500200-07-5A	0.5 – 2	32	1	0.7	2:1	2:1	5	295
JS2-00500200-15-5A	0.5 – 2	32	1	1.5	2:1	2:1	5	295
JS2-01000400-08-5A	1 – 4	27	1	0.8	2:1	2:1	5	200
JS2-01000400-20-5A	1 – 4	27	1	2	2:1	2:1	5	200
JS2-02000600-08-5A	2 – 6	22	1	0.8	2:1	2:1	5	125
JS2-02000600-20-5A	2 – 6	22	1	2	2:1	2:1	5	125
JS2-02000800-08-0A	2 – 8	22	1.25	0.8	2:1	2:1	0	125
JS2-02000800-20-0A	2 – 8	18	1.25	2	2:1	2:1	0	125
JS3-02001800-25-5A	2 – 18	23	1.8	2.5	2.5:1	2.5:1	5	150
JS3-02001800-50-5A	2 – 18	23	1.8	5	2.5:1	2.5:1	5	150
JS4-02001800-22-5A	2 – 18	30	2	2.2	2.5:1	2.5:1	5	200
JS4-02001800-50-5A	2 – 18	30	2	5	2.5:1	2.5:1	5	200
JS3-02002600-33-5A	2 – 26	21	2.5	3.3	2.5:1	2.5:1	5	150
JS3-02002600-40-5A	2 – 26	21	2.5	4	2.5:1	2.5:1	5	150
JS3-06001800-16-5A	6 – 18	23	1.8	1.6	2:1	2:1	5	125
JS3-06001800-30-5A	6 – 18	23	1.8	3	2:1	2:1	5	125
JS4-06001800-145-5A	6 – 18	31	2	1.45	2:1	2:1	5	200
JS4-06001800-30-5A	6 – 18	31	2	3	2:1	2:1	5	200

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18 to 40 GHz Design

MODEL NUMBER	FREQUENCY RANGE (GHz)	GAIN (dB, Min.)	GAIN VARIATION (±dB, Max.)	NOISE FIGURE (dB, Max.)	VSWR IN	VSWR OUT	POWER OUT @ 1 dB COMP. (dBm, Min.)	DC POWER @ +15 V (mA, Nom.)
MULTIOCTAVE BAND AMPLIFIERS (continued)								
JS3-08001800-16-5A	8 – 18	24	1.5	1.6	2:1	2:1	5	150
JS3-08001800-30-5A	8 – 18	24	1.5	3	2:1	2:1	5	150
JS4-08001800-145-5A	8 – 18	32	2	1.45	2:1	2:1	5	200
JS4-08001800-30-5A	8 – 18	32	2	3	2:1	2:1	5	200
JS3-12002600-25-5A	12 – 26	22	2.5	2.5	2.2:1	2.2:1	5	150
JS3-12002600-35-5A	12 – 26	22	2.5	3.5	2.2:1	2.2:1	5	150
JS4-12002600-22-5A	12 – 26	32	2.2	2.2	2:1	2:1	5	200
JS4-12002600-35-5A	12 – 26	32	2.2	3.5	2:1	2:1	5	200
JS3-18004000-38-5A	18 – 40	16	2.5	3.8	2.5:1	2.5:1	5	150
JS3-18004000-50-5A	18 – 40	16	2.5	5	2.5:1	2.5:1	5	150
JS4-18004000-30-5A	18 – 40	23	2.5	3	2.5:1	2.5:1	5	200
JS4-18004000-50-5A	18 – 40	23	2.5	5	2.5:1	2.5:1	5	200
ULTRAWIDE BAND AMPLIFIERS								
JS2-00100200-07-5A	0.1 – 2	32	1	0.7	2:1	2:1	5	295
JS2-00100200-15-5A	0.1 – 2	32	1	1.5	2:1	2:1	5	295
JS2-00100400-08-5A	0.1 – 4	27	1	0.8	2:1	2:1	5	200
JS2-00100400-12-5A	0.1 – 4	27	1	1.2	2:1	2:1	5	200
JS2-00100600-10-3A	0.1 – 6	23	1.5	1	2:1	2:1	3	175
JS2-00100600-20-3A	0.1 – 6	23	1.5	2	2:1	2:1	3	175
JS2-00100800-13-0A	0.1 – 8	20	1.5	1.3	2:1	2:1	0	175
JS2-00100800-25-0A	0.1 – 8	20	1.5	2.5	2:1	2:1	0	175
JS3-00101000-20-5A	0.1 – 10	23	1.5	2.0	2.5:1	2:1	5	150
JS3-00101000-35-5A	0.1 – 10	23	1.5	3.5	2.5:1	2:1	5	150
JS3-00101200-21-5A	0.1 – 12	23	1.5	2.1	2.5:1	2:1	5	150
JS3-00101200-35-5A	0.1 – 12	23	1.5	3.5	2.5:1	2:1	5	150
JS3-00101800-24-5A	0.1 – 18	23	1.8	2.4	2.5:1	2.2:1	5	150
JS3-00101800-40-5A	0.1 – 18	23	1.8	4	2.5:1	2.2:1	5	150
JS4-00101800-23-5A	0.1 – 18	29	1.8	2.3	2.5:1	2.2:1	5	200
JS4-00101800-40-5A	0.1 – 18	29	1.8	4	2.5:1	2.2:1	5	200
JS4-00102000-25-5A	0.1 – 20	28	1.8	2.5	2.5:1	2.5:1	5	200
JS4-00102000-35-5A	0.1 – 20	28	1.8	3.5	2.5:1	2.5:1	5	200
JS3-00102600-33-5A	0.1 – 26	20	2.5	3.3	2.5:1	2.5:1	5	150
JS3-00102600-42-5A	0.1 – 26	20	2.5	4.2	2.5:1	2.5:1	5	150
JS4-00102600-28-5A	0.1 – 26	27	2.5	2.8	2.5:1	2.5:1	5	200
JS4-00102600-50-5A	0.1 – 26	27	2.5	5	2.5:1	2.5:1	5	200
JS4-00104000-65-5A	0.1 – 40	14	4.5	6.5	2.75:1	2.75:1	5	200
JS4-00104000-85-5A	0.1 – 40	14	4.5	8.5	2.75:1	2.75:1	5	200

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AIA Calls For Structural Changes In Aerospace Industry

WASHINGTON, DC—John W. Douglass, President and CEO of the Aerospace Industries Association (AIA) says that converging forces in global politics and national security will make 2002 a pivotal year for aerospace. Douglass calls for change in three structural areas of aerospace: research-and-development (R&D) funding, Department of Defense (DoD) procurement, and export licensing, as ways to sustain growth in the industry.

"The events of September 11 and the war in Afghanistan could act as catalysts to ultimately strengthen aerospace and its role in national security and the economy."

At the 37th annual Year-End Review and Forecast Luncheon, Douglass told 300 media industry and government representatives that the events of September 11 and the war in Afghanistan could act as catalysts to ultimately strengthen aerospace and its role in national security and the economy. Douglass opines that the coming months will provide a unique opportunity for the Bush administration and Congress to reshape the aerospace industry and the environment in which it operates. He adds that the Commission on the Future of the US Aerospace Industry was primed to provide insight and leadership to strengthen and redirect the industry.

Douglass wants increased federal funding of R&D—particularly for Federal Aviation Administration (FAA) and National Aeronautics and Space Administration (NASA) civil aeronautics. He notes that although military R&D funding has set US military technology 10 to 15 years ahead of the rest of the world, there has been insufficient funding to develop and process new, affordable technologies to advance civil aviation systems. "Europe has pulled even with the US," Douglass says, adding that we need affordable generic technologies to improve civil aviation safety, security, system capacity, and noise and emission levels.

He also calls for increased funding for DoD procurement—especially for combat and support aircraft. "Aging US cargo, electronic-surveillance, and medical-evacuation fleets need to be replaced by commercial aircraft, modified for military purposes. DoD can save huge amounts of money by using existing commercial designs," he adds, "and there is no better time to begin that than in 2002."

He says that the third area, reform of export control licensing, and a major initiative for the AIA this year, would be necessary for better national—and international—security as well as increased US competitiveness worldwide.

Despite all the recent problems, Douglass says that there is reason for optimism. He states, "This administration, the Commission on the Future of the US Aerospace Industry, and the international coalition fighting terrorism are the right forces converging at the right time to fix aerospace problems presented by September 11, as well as lingering structural problems related to the transition away from the Cold War era."

Kudos

The National Academy of Television Arts and Science has presented the EMMY award, the most important award in the television world, to Rohde & Schwarz. The award went to the TV Test Transmitter SFQ for its performance features and its contribution to the further development of digital TV...At Northrop Grumman Electronic Systems Vendor Day in Baltimore, MD, Tyco Electronics' Global Industrial Commercial North America Business Unit was awarded the Bronze Award as part of Northrop Grumman's Preferred Supplier Program. Tyco Electronics was given the award due to its performance in the areas of quality, delivery, engineering capability, and support...The IMAPS Educational Foundation has been renamed in honor of Sidney J. Stein. The Stein Educational Foundation provides multiple $12,000 grants annually to support postgraduate research in electronics. This is the first time that $12,000 grants have been awarded. The grants had been $5000...Mimix Broadband, a developer of gallium-arsenide (GaAs) monolithic microwave integrated circuits (MMICs), has been registered to ISO 9001, the internationally recognized quality management system standard that is administered by the International Organization for Standardization...EMS Technologies, Inc. has announced that its subsidiary, LXE, Inc., has received a 2001 Mobile Star Award™ from MobileVillage, a mobile and wireless market development company. LXE won for best customer satisfaction in the mobile hardware category. A total of 16 winners were picked by subscribers of MobileVillage's weekly newsletter, *MobileVillage Weekly*. The annual Mobile Star Awards provide technology users and vendors with the opportunity to recognize the leaders in their industry in categories such as mobile leadership and best products. MRF

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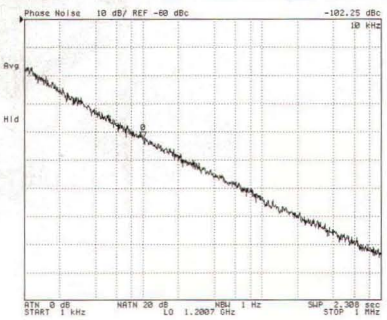
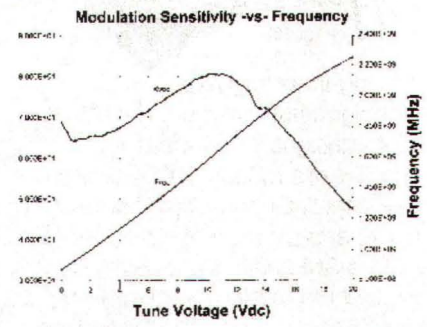
Part Number	Frequency Range(MHz)	Tuning Voltage	Typical 10 kHz Phase Noise	Supply Voltage	Output Power	Package Size
VCO790-600T	400-800	0.0 - 20.0	-102 dBc/Hz	+5 V	+3 dBm	0.5 x 0.5 x 0.18 in.
VCO790-1500T	1000-2000	0.0 - 20.0	-98 dBc/Hz	+5 V	+2 dBm	0.5 x 0.5 x 0.18 in.
VCO790-2300T	2100-2500	1.0 - 4.0	-89 dBc/Hz	+5 V	+3 dBm	0.5 x 0.5 x 0.18 in.
VCO793-600T	400-800	0.0 - 20.0	-104 dBc/Hz	+12 V	+7 dBm	0.5 x 0.5 x 0.18 in.
VCO793-1500T	**1000-2000**	**0.0 - 20.0**	**-99 dBc/Hz**	**+12 V**	**+7 dBm**	**0.5 x 0.5 x 0.18 in.**

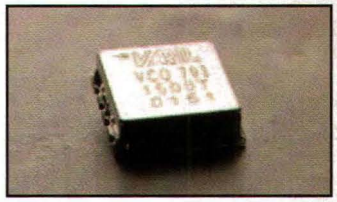
Actual data for VCO793-1500T

Phase noise from HP3852 for 1000-2000 MHz VCO



Tuning Sensitivity from HP3852 for 1000-2000 MHz VCO





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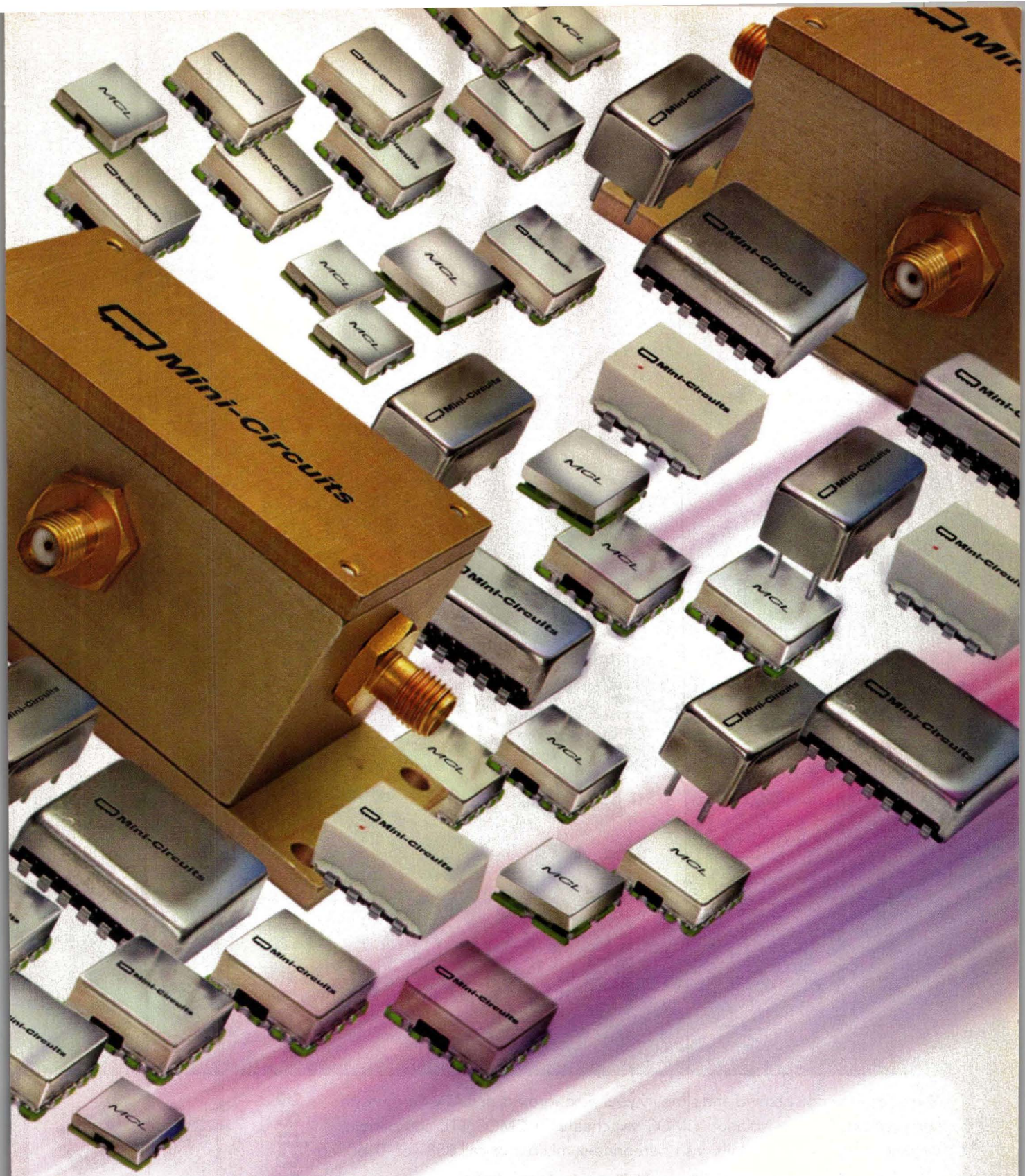

[clear communications ICs]

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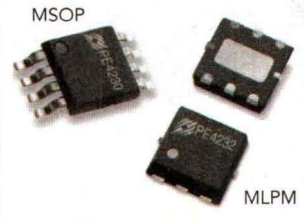
Product	Frequency (MHz)	IIP3[a]/IIP2[b] (dBm, 1 GHz)	P1dB (dBm, 1 GHz)	Ins Loss (dB, 1 GHz)	Isolation (dB, 1 GHz)	Package
Reflective SPDT RF Switches						
PE4210	10–3000	35[a]	15	0.40	36	8L MSOP
PE4220	10–2500	39	21	0.25	36	8L MSOP
PE4230	10–3000	50	30	0.44	38	8L MSOP
PE4235	10–4000	35	15	0.42	42	6L MLPM
PE4237	10–4000	50	30	0.46	44	6L MLPM
CATV High Isolation Absorptive SPST Switches						
PE4232	5–1300	80[b]	30	0.75	53	6L MLPM
PE4240	5–1300	80	30	0.75	47	6L MLPM

Switches



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NEWS

Taking On The Recession With "Breakthrough" Products

Aggressive product introductions through the use of multidiscipline product-development teams are one approach to energizing a lackluster economy.

Challenging times call for innovative solutions. And rather than complain about a difficult economy, management at Anaren Microwave, Inc. (Syracuse, NY) has chosen an aggressive strategy of multiple-product introductions to spur business growth. Dubbed "Breakthrough 2002," the mix of new products includes several which have been unavailable in these form factors or package styles until now, including a surface-mount circulator as well as multilayer vector modulators and delay lines.

JACK BROWNE
Publisher/Editor

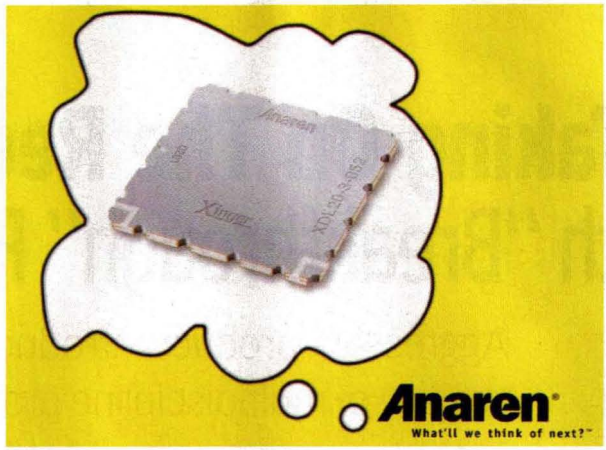
Despite the recent loss of its founder and former CEO, Hugh Hair **(see sidebar, p. 38)**, the company is healthy and well-poised for growth. The multiple product introductions are planned not as a one-time event, but on a regular schedule throughout the year as part of the company's TeamNEXT initiative in which cross-discipline teams (such as mechanical engineers and electrical engineers) work on the development of new products according to input from customers and marketing specialists. According to Larry Sala, President and CEO of Anaren, "Infrastructure and service providers in our industries are expected to introduce performance and cost breakthroughs on a regular basis. With BreakThrough 2002, we're just doing our part on the component side, and, we believe, establishing what will be the norm for players at our tier for years to come." He notes that some of the products represent incremental improvements, but that the development program is also intended to create innovations rather than wait for an accidental discovery. "We have to introduce leapfrogging technologies like these on a regular, planned basis," he says.

1. The X180L-100 is a true surface-mount left-hand-polarized circulator for DCS base-station transmit applications from 1805 to 1880 MHz.

One of the most exciting products is a Xinger®-brand circulator—the first true surface-mount circulator. Traditionally, circulators have been limited in miniaturization by the need to mount a ferrite sphere in a resonant cavity. The new Xinger circulators take advantage of the multilayer circuit technology employed in the company's lines of Xinger hybrid couplers to create leadless surface-mount circulators that measure only 1.0 in. (2.5 cm) on a side and are only 0.29 in. (0.74 cm) thick. According to Sala, "Quite deliberately, we've wedded the compact, surface-mount concept of our Xinger products to the ferrite discipline. We saw the whole arena of circulators and isolators as ripe with potential for major

reductions in cost and labor at the OEM side."

The first three circulator models **(Fig. 1)** are the X180x-100, the X190x-100, and the X210x-100, with respective frequency ranges of 1805 to 1880 MHz [for digital-communications-services (DCS) base-station transmit applications], 1930 to 1990 MHz [for personal-communications-services (PCS) base-station transmit applications], and 2110 to 2170 MHz [for Universal Mobile Telecommunications Services (UMTS) and other third-generation (3G) base-station transmit applications]. (These models are available in left-hand and right-hand polarized versions, for example, as models X180L-100 and X180R-100, respectively.) Despite their miniature size, these circulators can handle average power levels of 100 W continuous wave (CW) due in large part to almost negligible insertion loss of 0.29 dB. The small size does not cause a penalty in isolation, however, since these circulators boast 23-to-25-dB isolation at temperatures from −20 to +85°C. The return loss is at least 23 dB for all models. The circulators are non-lead (Pb) solder paste compatible for use in European countries where Pb contamination in electronic components is a key concern. According to Product Manager H.P. Ostergaard, the manufacturing process produces superb uniformity across lots, with "excellent part-to-part and batch-to-batch repeatability for increased yield. We can provide even better performance when these circulators are specified in combination with matching terminations or attenuators from our RF Power-brand family."



2. The XDL20-3-050 is a surface-mount delay line capable of 5-ns delay time at PCS frequencies from 1930 to 1990 MHz. It measures only 1.0 × 1.0 × 0.44 in. (2.54 × 2.54 × 1.18 cm).

The new way of thinking was also applied to delay lines, which are usually physically large circuits or coaxial cables that achieve the electrical delays needed for some multichannel receivers (Rxs) as well as in predistortion circuits and feedforward amplifiers. These first-ever surface-mount delay lines are available for applications through 3 GHz. They measure only 1.0 in. (2.5 cm) on a side and 0.05 in. (0.13 cm) thick (for 2 and 5 ns delays) and only 0.1 in. (0.3 cm) thick (for 10 ns delay), but can provide the equivalent electrical delays of much larger sections of semirigid coaxial cable. They do not suffer the phase deviations induced by changes in temperature that characterize coaxial-cable delay lines. "Like the circulators, these delay lines offer excellent part-to-part repeatability," notes H.P. Ostergaard, "and without the production-line tuning needed to achieve repeatability with coaxial-cable delay lines."

Ostergaard adds that delay-line components have often been the cause of delays in their customers' production lines. "When we were exploring the downsides of traditional delays, we couldn't help but find it ironic that installing coax delays caused so many delays in production time," he says. The surface-mount delay lines, which are available in tape-and-reel packag-

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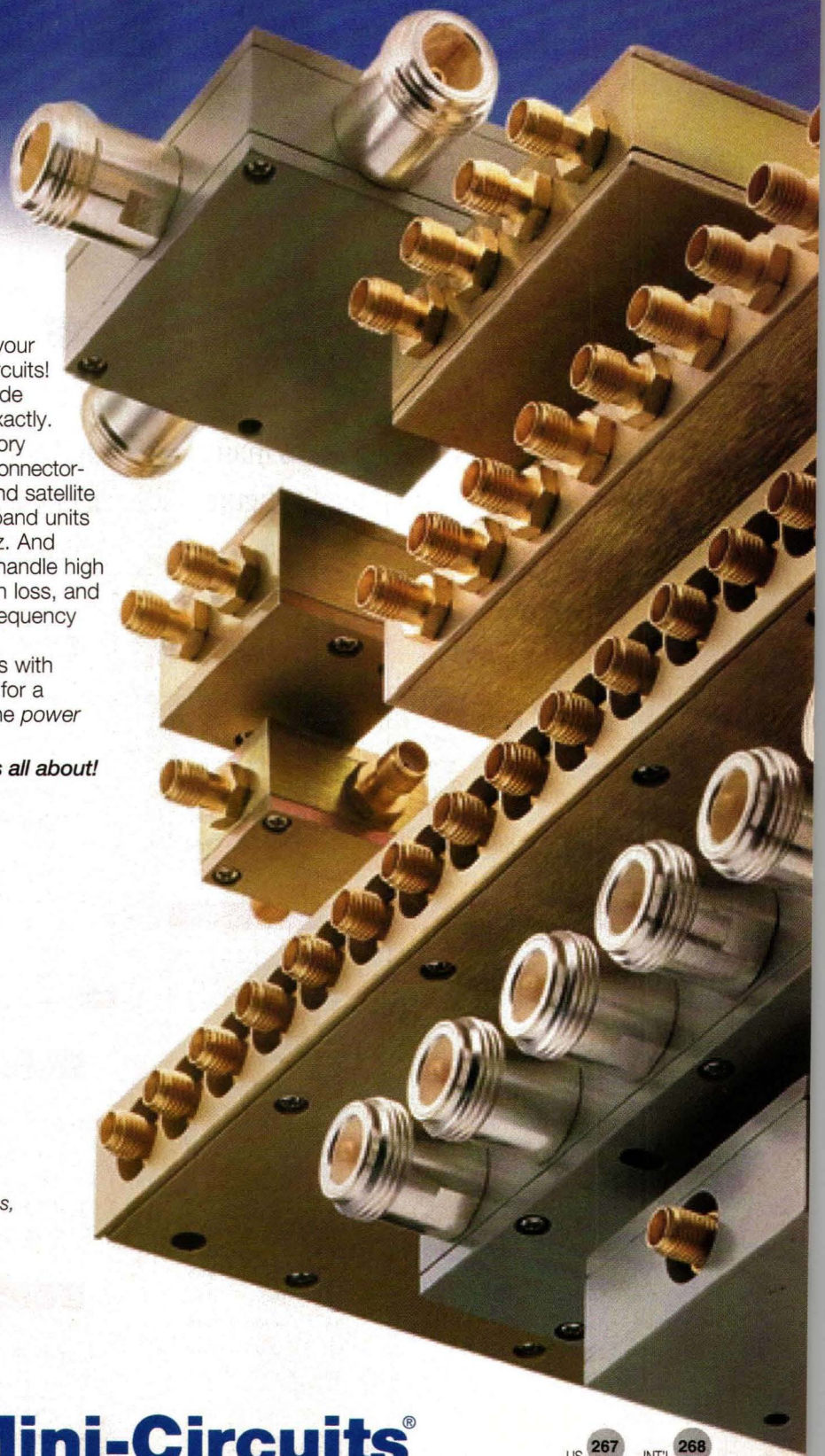

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Model	No. Of Models	Freq. Range (GHz)
2WAY-0°	151	0.50-10.0
2WAY-90°	4	1.00-4.20
2WAY-180°	2	1.00-2.49
3WAY	41	0.50-4.20
4WAY	83	0.47-8.40
5WAY	6	0.80-1.98
6WAY	25	0.80-5.00
7WAY	4	1.00-1.99
8WAY	62	0.50-8.40
9WAY	3	0.80-4.80
10WAY	11	0.75-2.40
12WAY	9	0.50-4.20
14WAY	1	0.90-0.99
16WAY	28	0.47-4.80
32WAY	1	0.95-1.75

For detailed model numbers, specifications, and prices, consult our web site, RF/IF Designer's Guide, or call Mini-Circuits.

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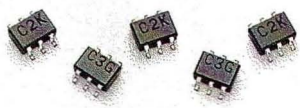

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3V Wideband RFIC Amplifiers

Part Number	Function	3dB BW (MHz)	Gain (dB)	P_{SAT} (dBm)	NF (dB)	I_{CC} (mA)	Test Freq (MHz)
UPC2745TB	General Purpose	2700	12	-1	6	7.5	500
UPC2746TB	General Purpose	1500	19	0	4	7.5	500
UPC2747TB	Low Noise - 1900 MHz	1800	12	-7	3.3	5	900
UPC2748TB	Low Noise - 1900 MHz	1500	19	-3.5	2.8	6	900
UPC2749TB*	Low Noise - 900 MHz	2900	16	-6	4	6	1900

*For GPS RF amplifier applications

3V VCO Buffer Amplifiers

Part Number	3dB BW (MHz)	Gain (dB)	P_{1dB} (dBm)	NF (dB)	I_{CC} (mA)	Test Freq (MHz)	Package Style
UPC8128TB	1900	12	-2	—	- 2.8	900	S06
UPC8151TB	1900	12	4	—	4.5	900	S06
UPC8152TB	1900	23	-5	—	5.6	900	S06
NEW UPC8178TB*	>2400	11.2	-4	5.5	1.9	1000	S06
NEW UPC8179TB*	>2400	13.5	3	5	4	1000	S06

*Specified for operation at frequencies to 2.4 GHz

3V RF Driver Amplifiers

Part Number	3dB BW (MHz)	Gain (dB)	P_{SAT} (dBm)	NF (dB)	I_{CC} (mA)	Test Freq (MHz)	Package Style
UPC2762TB	2900	14.5	8.5	6	27	1900	S06
UPC2763TB	2400	20	8	5	27	1900	S06
UPC2771TB	2400	20	13	5	35	900	S06
NEW UPC8181TB	4000	19	9.5	4.5	23	900	S06
NEW UPC8182TB	2900	21.5	10.5	4.5	30	900	S06

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5V Wireless & Multimedia Amplifiers

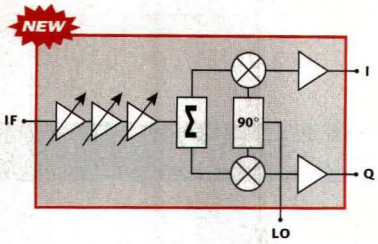
Part Number	3dB BW (MHz)	Gain (dB)	P_{SAT} (dBm)	NF (dB)	I_{CC} (mA)	Test Freq (MHz)	Packag Style
UPC1678GV	1900	23	17.5	6	49	500	S08
UPC2791TB	1900	12	4	5.5	17	500	S06
UPC2792TB	1200	22	5	4.5	19	500	S06
UPC2708TB	2900	15	10	6.5	26	1000	S06
UPC2709TB	2300	23	11.5	5.0	25	1000	S06
UPC2710TB	1000	33	13.5	3.5	22	500	S06
UPC2711TB	2900	13	1	5	12	1000	S06
UPC2712TB	2600	20	3	4.5	12	1000	S06
UPC2713T	1200	29	7	3.2	12	500	T06
UPC2776TB	2700	23	8.5	6	25	1000	S06
NEW UPC3215TB	2900	20.5	3.5	2.3	14	1500	S06

3V Silicon Downconverters...

Part Number	RF Freq (MHz)	IF Freq (MHz)	CG (dB)	IIP_3 (dBm)	NF (dB)	I_{CC} (mA)	Test Fre (MHz)
UPC2756TB	100-2000	10-300	14	-14	13	5.9	900
UPC2757TB	100-2000	20-300	13	-15	13	5.6	800
UPC2758TB	100-2000	20-300	17	-14	13	11	800
UPC8112TB	800-2000	100-300	13	-10	11	8.5	900

... and Upconverters

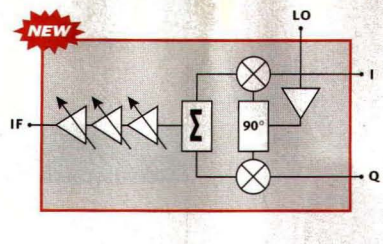
Part Number	RF Freq (MHz)	IF Freq (MHz)	CG (dB)	IIP_3 (dBm)	NF (dB)	I_{CC} (mA)	Test Fre (MHz)
UPC8106TB	100-2000	50-400	9	-4.5	8.5	9	900
UPC8163TB	100-2000	50-400	9	+0.5	12.5	16.5	900
NEW UPC8172TB	800-2500	50-400	8.5	-2	9.5	9	190
NEW UPC8187TB	800-2500	50-400	11	-1	9	15	190



AGC + I/Q Demodulators for W-CDMA Tx/Rx

UPC8190K
UPC8194K

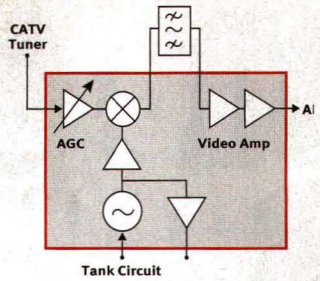
- 3V Operation @ 9mA
- IF Freq. Plan @ 180 or 380 MHz
- 77dB Gain/97dB AGC Range
- +3dBm IIP_3 @ −10dB Gain
- 10MHz IQ Bandwidth
- 20 Pin QFN package



AGC + I/Q Modulators for W-CDMA Tx/Rx

UPC8191K
UPC8195K

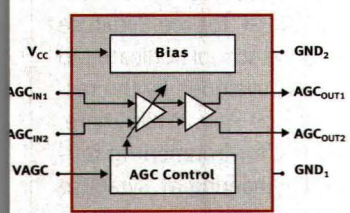
- 3V Operation @ 9mA
- IF Freq. Plan @ 380 or 570 MHz
- −10db Output Power
- 95dB AGC Range
- 10MHz IQ Bandwidth
- 20 Pin QFN package



Cable Modem/Settop IC

UPC2798GR

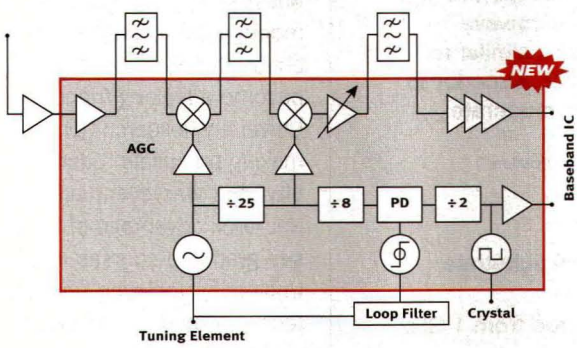
- On-Chip VCO and Downconversion Mixer
- 30dB AGC Range
- 3 V_{PP} Video Amp Output
- V_{CC} = 9V

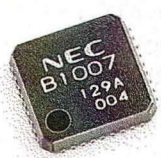


Differential AGC Amplifiers

UPC3217GV
UPC3218GV
UPC3219GV

- 60dB Dynamic Range
- 3217 = 50dB gain
  3218 = 55dB gain
  3219 = 42dB gain
- IM3 < -55dBc
- 0.5−2.5V Control Voltage
- 1.25 V_{PP} @ Z_L=240Ω
- 8 pin SSOP





Integrated GPS Receiver IC

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- 4MHz Baseband Output
- 36 pin QFN Package

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NEWS

Anaren Loses Founder

JACK BROWNE
Publisher/Editor

Hugh Hair, founder and former CEO of Anaren Microwave, Inc. (East Syracuse, NY) passed away on January 27th, leaving behind a wife, family, friends, the 300-plus employees of a 35-year-old high-frequency-electronics firm, and many members of this industry who shall miss him. Hair had been CEO of the firm as recently as 1997, and had only stepped down as Chairman of the Board last November. He founded Anaren Microwave in 1967 with long-time friend and current Anaren CTO Carl Gerst, Jr.

Hugh Hair was recruited by General Electric from Glasgow University in 1960, moving from Scotland with his wife Renee to the Syracuse, NY area. He had earned Bachelor of Science and Master of Science degrees in Electrical Engineering from the school, and applied his knowledge to the development of phase shifters, phased-array radar, missile-site radar, and the modulation scanned-array radar (MOSAR) for the defense contractor. He met Carl Gerst at GE and started what would become a lifelong friendship. Gerst left GE for the Syracuse University Research Corp., an arm of the school dedicated to making practical products of theories developed by the faculty, and was eventually joined there by Hair. While there, Hair took advantage of the educational environment to pursue graduate courses in Electrical Engineering at Syracuse University, focusing on sensory communications. It was a technical area that he would put aside with the founding and growth of Anaren, but would return to 30 years later.

While at the Syracuse University Research Corp., Hair developed the design of a system to detect incoming missiles for the US Navy. Although the military wanted the system put into production, Syracuse University Research Corp. was only chartered as a research organization. In 1967, Hair and Gerst decided to leave the firm to start Anaren and produce the system. Originally called Micronetics, the name of the company was changed due to a conflict with a California company also known as Micronetics. Anaren grew steadily as a defense contractor until defense spending slowed in the early 1990s. Based on the advice of young engineer Larry Sala, a boyhood friend of Hair's son Kevin, the company added commercial components for wireless applications to its mix of products and continued its steady growth through the present. Sala would go on to succeed Hair, and is now President and CEO of Anaren.

Hugh Hair retired as Anaren's CEO in 1997. He stepped down as Chairman of the Board in November 2001 to pursue his quest to help blind children see: Based on research of the sonar-like techniques of bats, he had hoped to develop a way to use sound to help the blind "see" the dimensions of a room and detect the location of objects within it.

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3 year full warranty

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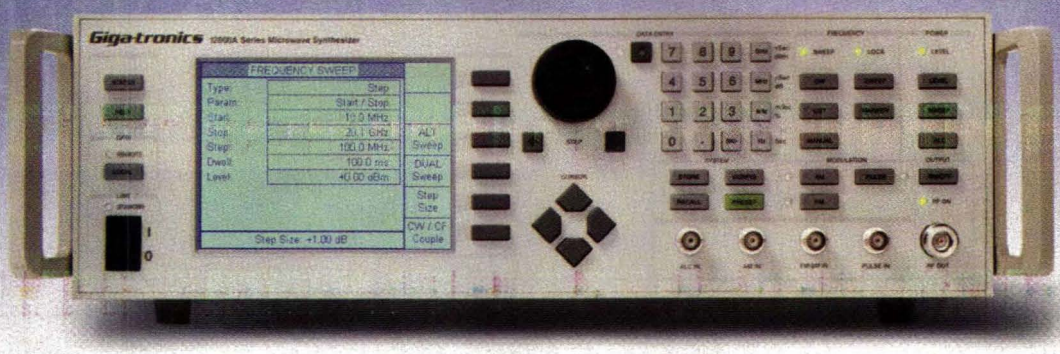

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Get to know the 12000A Microwave Synthesizer at **www.gigatronics.com**

Good news for microwave engineers working up to 20 GHz. With a frequency switching speed of 500µs, the 12000A outperforms the competition for a fraction of the price. Take advantage of 15 dBm standard power and legendary spectral purity. For satellite communication and Ku band links, 20 GHz coverage, high power and fast frequency switching are an immediate benefit. Plus, you can test systems and components used in these systems. For wireless local loop, the 8 GHz models provide the necessary frequency range without making you pay for the full 20 GHz microwave spectrum. And for fixed wireless systems, the 12000A is well-equipped to accurately test systems and components. In fact, when characterizing the frequency response of these system, fast switching assures minimum test time for maximum profit. 12000A application notes and more are yours for the taking at **www.gigatronics.com**.

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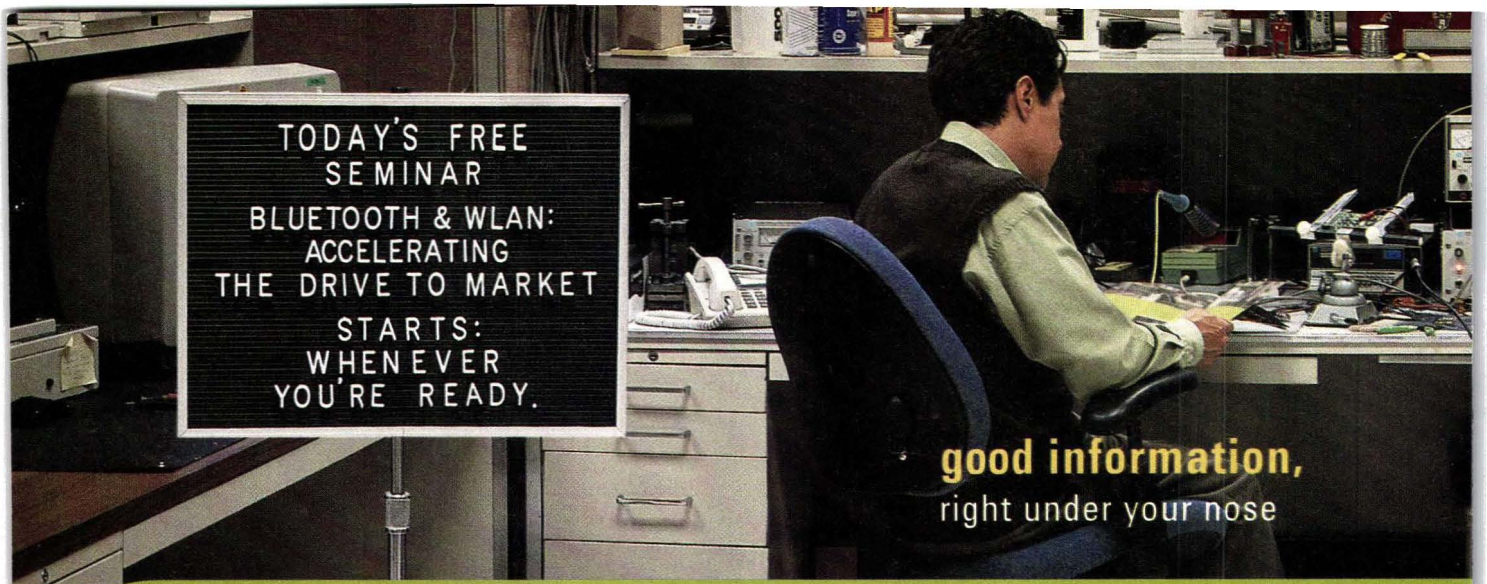


good information,
right under your nose

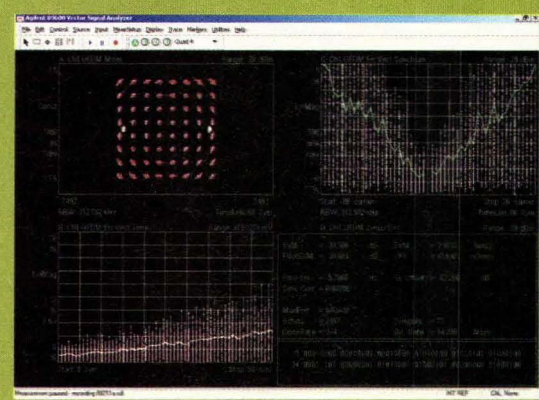
The original idea was simple: use wireless links to replace the tangle of cables that connect PCs, PDAs, mobile phones and more. Of course, turning that idea into reality—without much time for analysis—has been anything but simple. Perhaps we can help.

As you work to create the *Bluetooth* and Wi-Fi 5 products of today and tomorrow, we're busy working with standards committees and engineers like you to address challenges such as interoperability, certification and quality. Here are three quick samples of the insights we've gathered along the way.

Ensuring interoperability.

Many people attribute the popularity of Wi-Fi devices to WECA testing that certifies the interoperability of products from multiple vendors. Of course, the roots of interoperability reach back to the early stages of product development when each manufacturer (or silicon supplier) adds value by optimizing its designs in unique ways. It's a good practice, but one that may leave your products working well together but not with devices from another maker.

Developers tell us interoperability is often a matter of tweaking a transmitter or receiver design. For transmitters, error vector magnitude (EVM) versus time or channel is a measure of modulation quality that can highlight underlying problems such as nonlinear distortion, phase noise and spurious signals. Conversely, one way to make receivers more forgiving of nonideal transmitters is to test them with a variety of impaired signals—in hardware with a flexible signal-generation solution, in a computer-based simulation, or in a system that links both methods.



For this IEEE 802.11a signal, the overall EVM measurement is acceptable but viewing EVM versus time (lower left) and channel (upper right) shows the effect of a timing error.

Clearing the qualification hurdle.

Getting through certification and regulatory testing quickly is obviously a major challenge. With WLAN, it's often useful to focus pre-qualification work on the "PHY" (RF) layer because certification tests it indirectly. Digitized recording simplifies the analysis of problem signals, and replaying captured transmissions from other devices enables repeatable PHY tests.

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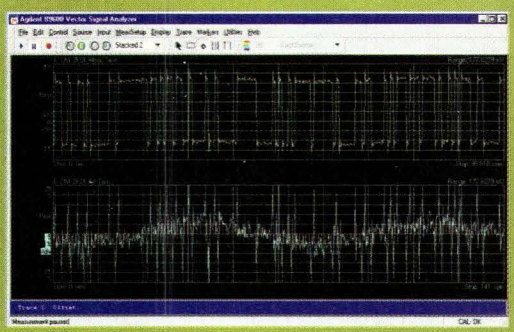
To achieve *Bluetooth* qualification, your product must pass all 16 sections of the RF test specification. From what we've seen, test results are as unique as every design and no single test yields consistent failures. However, most receivers exceed the sensitivity specification so the nine transmitter tests tend to be the biggest concern.

Problems are often specific to the type of transmitter. In FM-based designs, frequency drift is a common issue. Digital noise on the power supply can affect modulators and VCOs, producing ripple in a frequency drift measurement. For IQ-based modulators, the culprit is often modulation quality. They're not called out in the test specification, but measurements such as FSK error, magnitude error and the eye diagram help identify modulation quality problems.



The FSK error display can highlight the effects of unwanted frequency modulation, which may indicate the presence of spurious signals in the modulator.

Optimizing manufacturing test.

Although many Wi-Fi and *Bluetooth* products are designed for consumer applications, most are complex enough to warrant some level of testing on the manufacturing line. But how much testing?

Combining your device expertise with our instrument knowledge, we can create an optimized test program that needs only a subset of the relevant RF test specs. If common test modes are designed in, it's also possible to accelerate some of the tests. For OEMs who purchase and integrate *Bluetooth* subsystems, testing can focus on the PHY layer rather than the protocol layer.

Getting the rest of the story.

Sharing insights and best practices is just one way Agilent can help accelerate the drive to market with new wireless networking products. The Agilent Interoperability Certification Labs and Agilent's network of test partners are ready to help, too: they've tested hundreds of Wi-Fi devices and can offer vital insights into clearing the qualification hurdle.

To learn more, please visit www.agilent.com/find/wn, where you can request a FREE CD-ROM packed with articles, solution guides, and application notes such as "RF Testing of Wireless LAN Products" and "Verifying *Bluetooth* Baseband Signals."

Agilent Technologies

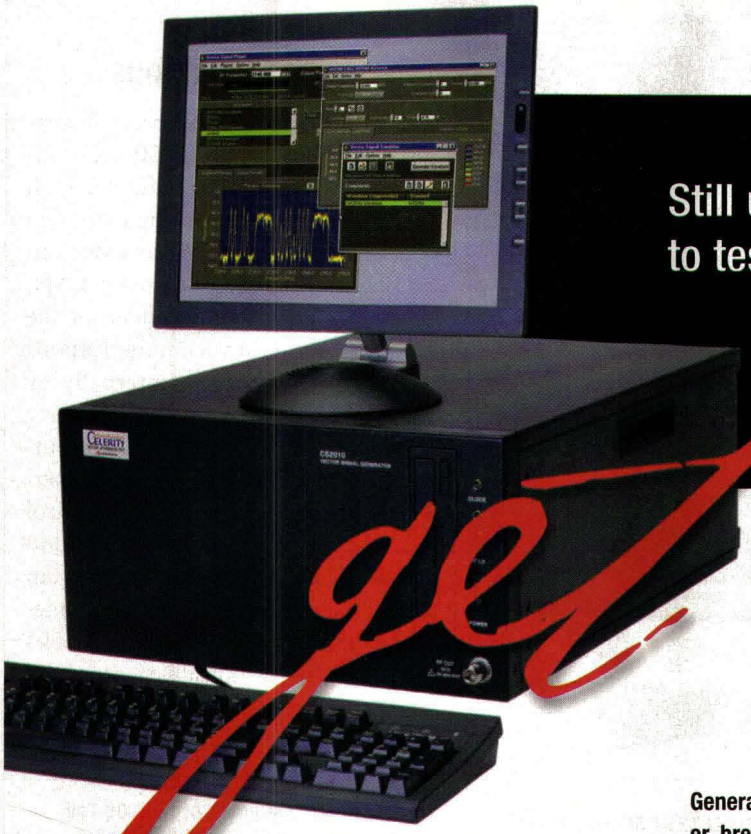
dreams made real

ing for use with automated assembly equipment, are meant to eliminate production delays due to the use of labor-intensive coaxial delay lines.

Used in the low-power loop of feedforward amplifiers, standard delay times for Anaren's new surface-mount components are 2, 5, and 10 ns; other delay times are available upon request. Examples of new models include the XDL20-3-050 **(Fig. 2)** and the XDL20-6-100. Both can be used from 400 to 3500 MHz. The former provides 3 ±0.03 ns delay in the Advanced Mobile Phone Service (AMPS) band from 869 to 894 MHz, with 5 ±0.05 ns delay in the PCS band from 1930 to 1990 MHz. It achieves 0.1-dB peak-to-peak amplitude flatness in both bands, with at least 20-dB return loss and better than 0.55-dB/ns insertion loss. The delay line handles 80-W CW power at cellular frequencies and 40-W CW power at PCS frequencies. The model XDL20-6-100 provides longer delay times of 5.80 ±0.06 ns in the AMPS band and 9.70 ±0.10 ns in the PCS band.

Perhaps as impressive as either of these two product types is the model XVM190A Xinger vector modulator. Based on the same multilayered-softboard technology as the circulators and delay lines, the XVM190A is a true surface-mount vector modulator that, unlike the circulators and delay lines, incorporates active and passive components. The modulator has an RF input port, ports for in-phase (I) and quadrature (Q) modulation signals, and an RF output port. The component, which measures only 1.00 × 0.65 × 0.11 in. (2.54 × 1.65 × 0.25 cm), provides a full 360-deg. phase control for PCS (1930 to 1990 MHz) linearization applications. It can be used for predistortion and for the vector-correction circuitry in the main or error loops of a feedforward amplifier. The tiny vector modulator has maximum insertion loss of 9 dB with input VSWR of 1.22:1 and output VSWR of 1.42:1. The deviation from linear phase is no worse than ±1 deg. at the minimum attenuation state and no worse than ±3 deg. at the maximum attenuation (20-dB) setting.

In addition to the three product families mentioned, Anaren also announced new miniature Xinger-brand 3-dB couplers for coverage of 1700 to 2000 MHz, 1800 to 2000 MHz, 2000 to 2700 MHz, and 2500 to 2700 MHz bands, as well as the RF Power brand alumina [an alternative to beryllium oxide (BeO) and aluminum nitride (AlN)] resistors and terminations for applications up to 40 W. For more information on the "Breakthrough 2002" and other products, contact: **Anaren Microwave, Inc., 6635 Kirkville Rd., East Syracuse, NY 13057; (315) 432-8909, FAX: (315) 432-9121, Internet: www.anaren.com.** MRF

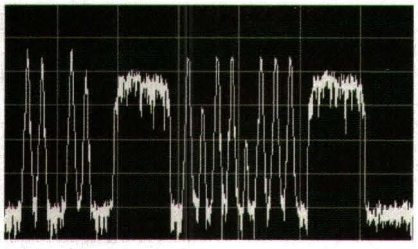


Still using multiple signal generators to test multi-carrier devices?

get real.

Generating multi-carrier test signals shouldn't try your patience or break your budget. Wrestling with multiple signal generators wastes precious test resources while sacrificing signal accuracy and realism. Why endure that expense and hassle when Celerity's new CS2010 Vector Signal Generator can create all the signals you need quickly and easily?

A single CS2010 VSG can generate up to six simultaneous WCDMA carriers, or virtually any combination of standard and custom signals within its super-wide 30 MHz signal bandwidth. Powerful, point-and-click software puts any signal at your command. An ultra-linear DAC, "Direct-to-IF" signal generation and careful up-converter design combine to eliminate the I/Q modulator errors found in traditional signal generators. Resulting signals possess unsurpassed precision, dynamic range and realism. You're assured of reliable, repeatable test results each and every time. If you need multi-carrier signals but want to skip all the frustration, you need a CS2010 VSG.



Actual multi-standard signal (GSM+EDGE+WCDMA) generated by the CS2010 VSG.

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Check out the CS2010 VSG and more in Orlando at CTIA, booth #1787.

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Analyzers monitor remote base stations

THE MODEL VC900 vector spectrum analyzer covers the 800-to-1000-MHz cellular bands and can be configured with three inputs for monitoring three sectors of a base station. The analyzer can be operated remotely through LAN, telephone or wireless modem, or the Internet. All of these communications methods are supported internally by the VC900 with no other third-party hardware or software. Virtual front-panel software provides a remote operator with complete access and control of all the functions of the VC900. Additionally, any number of operators can access the same instrument simultaneously. Optional logging software enables the VC900 to periodically measure and store parameters of interest. The unit measures 6.75 × 9.75 × 3.13 in. (17.15 × 24.77 × 7.95 cm).

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Tyco Reports First-Quarter Gain

TYCO INTERNATIONAL bucked the recent trend of poor earnings announcements by reporting outstanding figures for its first quarter of 2002 ending December 31, 2001. The company, with strong holdings in electronics markets that include M/A-COM, reported that diluted earnings per share for the first quarter were 74 cents, which represents a 17 percent increase over the pro forma earnings of 63 cents per share for the same quarter of the previous year.

The results for the first quarter have been restated to reflect the adoption of Staff Accounting Bulletin No. 101, and exclude good will amortization as a result of the adoption of Statement of Financial Accounting Standards No. 142. Revenues for the first quarter ending December 31, 2001 rose 25 percent to $10.1 billion compared to revenues of $8.0 billion for the first quarter of the previous year.

The company also reported that its earnings per share guidance for fiscal 2002 is $0.80 to $0.82 per share. This earnings estimate did reflect continued uncertainty about the near-term outlook for the electronics end markets. The company, which is divided into Electronics, Healthcare and Specialty Products, and Fire and Security segments, showed decreased revenue in the communication, telecommunications, printed circuit, as well as computer and consumer electronics end markets. There was also a 23-percent decrease at Tyco's Telecommunications unit. The company's international health-care business and the company's fire and security businesses, however, have contributed to the company's recent success.

Despite the optimism surrounding the positive first-quarter results, there was some concern on the part of investors regarding accounting practices at Tyco International, with the per share price dropping $6.96, or 19 percent, to $29.90 this past February 4th. The poor showing on Wall Street was linked to reports that Tyco had spent approximately $8 billion over the last three fiscal years on more than 700 acquisitions that were never announced to the public. Some financial reports pointed to similarities in accounting methods between Tyco International and the recently bankrupt energy giant Enron. MRF

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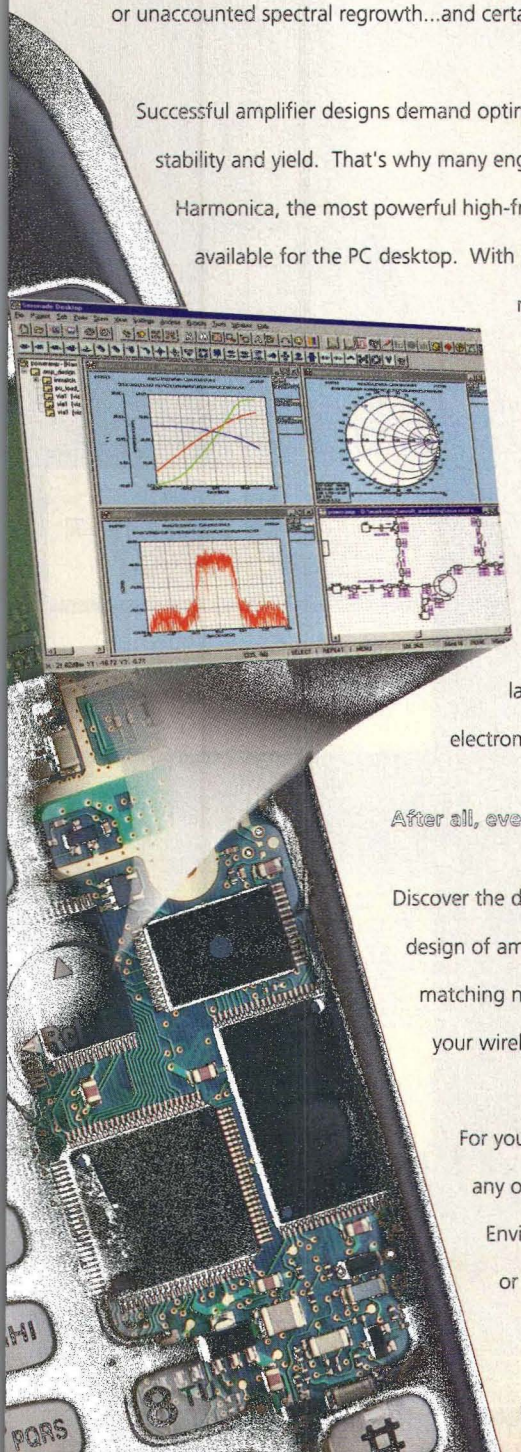

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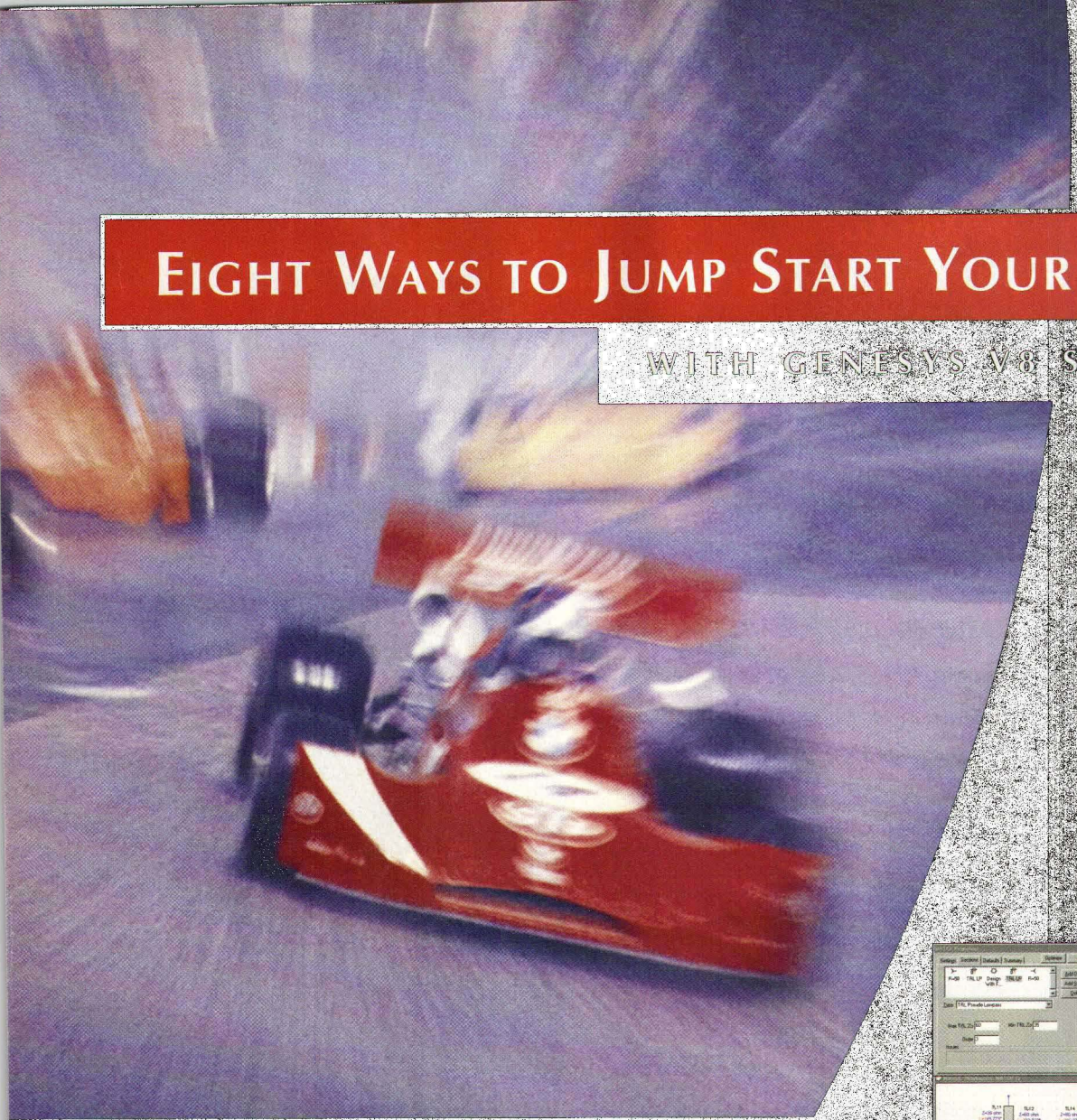

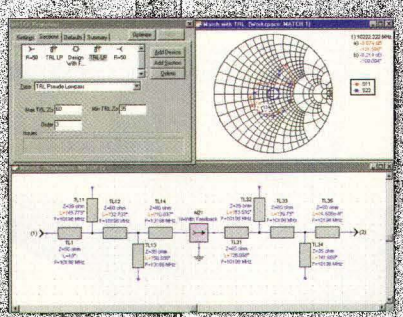

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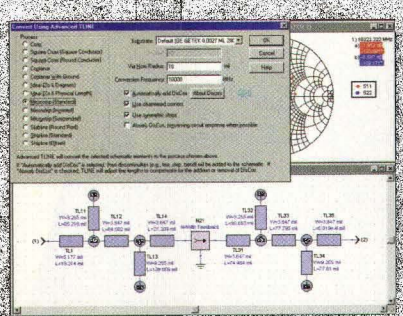

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CONTRACTS

Harris Corp.—Has signed a contract valued at approximately $200 million with General Dynamics United Kingdom Ltd. to supply the high-frequency subsystem for the United Kingdom's Bowman Tactical Radio Programme. Harris was selected as the preferred supplier in July 2001. Under the terms of the contract, Harris is responsible for providing approximately 10,000 high-frequency radio systems to UK Defense Forces. Harris' Bowman solution is based on the company's Falcon™ II Tactical Radio product family, which is in standardized use by US, NATO, Partnership for Peace (PfP), and other defense forces around the world. Local UK production of the high-frequency system will accelerate the growing standardization of the Falcon II radios worldwide. Deliveries of high-frequency radio systems will start during the next fiscal year and are expected to stretch over a five-year period.

EMS Technologies, Inc.—Announced that TRW, Inc. has selected EMS for a contract valued at approximately $22 million over approximately three years to deliver the Beam Forming Network (BFN) for the first two satellites of the AEHF (Advanced Extremely High Frequency) AEHF Project, the US Department of Defense (DoD) next-generation secure communications-satellite constellation. The BFN is a major component of the nuller subsystem, which provides anti-jam capabilities, a key function of the AEHF network that enables all branches of the US military to communicate in real time around the globe. The Air Force has expressed interest in purchasing three or more payloads to complete the constellation.

Sprint—Has been awarded a multiyear, multimillion-dollar agreement by Countrywide Credit Industries, Inc. for data and professional services. Countrywide also renewed its long-standing arrangement with Sprint for voice services at its call centers. As part of the expanded relationship, Sprint will double the bandwidth at more than 500 of Countrywide's branch sites and link them to core facilities in California and Texas. In addition, professional service consultants from Sprint E|Solutions will manage the project and provide hardware-implementation services.

Yokogawa Corp. of America—Signed a contract with The Gas Company to supply a turn-key distributed control system (DCS), field instruments, and electrical installation services for The Gas Company's synthetic natural gas plant in Honolulu, Hawaii on the island of Oahu. The Gas Company is using Yokogawa technology and turnkey project services to convert from a mainly pneumatic process-control plant to a fully integrated DCS control system.

FRESH STARTS

Anritsu Co., GGB Industries, and Keithley Instruments—Have formed a technology partnership resulting in the development of a test solution for wafer-level process monitoring of communications devices and other high-speed digital devices. Anritsu's vector-network-analyzer (VNA) technology and GGB's DC/RF probe cards are being integrated into Keithley's Model S400DC/RF ATE System, which help make it the first integrated system that can execute DC and RF tests independently and with a single probe touchdown.

RF Micro Devices, Inc.—Announced that it has agreed to purchase IBM's GPS development operation. The acquisition provides RFMD with advanced GPS technology and access to IBM's chip-scale packaging technology. The GPS development operation was the first to introduce GPS solutions using SiGe, which reduces size, power consumption, and noise figure, and enables higher levels of integration. As part of the transaction, IBM has agreed to transfer intellectual property associated with these products to RFMD, for which RFMD plans to file for patent protection.

Zetex—Widened its franchise agreement with Future Electronics to include The Americas and Europe. Future Electronics previously supported the Zetex range of power- and signal-management products in North America only, an arrangement that followed Future's acquisition of Zetex's incumbent US distributor Advent in 1998.

Gabriel Electronics—Acquired Endwave Corp.'s antenna business. The assets acquired in this transaction include intellectual property related to microwave and millimeter-wave antenna solutions such as the GemFire™ and WavShapr™ products, including patents and patent applications. Other assets include laboratory equipment and software systems used in design and development.

Renaissance Electronics Corp.—Signed a manufacturer's representative agreement with Blackhart, Inc. of Roy, UT. Under the terms of the agreement, Blackhart, Inc. will market Renaissance's products in Colorado, Utah, Idaho, and Montana.

Radiant Networks plc and STAR 21 NETWORKS AG—Have signed an agreement under which STAR 21 will trial Radiant's MESH*WORKS*™ architecture in Germany. This will be the first deployment of Radiant's Mesh technology in Europe. The trial will be a key component in STAR 21's initiative to develop and deploy broadband networks, using the most efficient enabling technologies. Radiant will deliver a fully supported two-phase implementation, which will be deployed and evaluated jointly by both companies. An initial pilot system is currently being deployed in Frankfurt, Germany, followed by a second-phase operation.

REMEC, Inc.—Has purchased, in a sale approved by the US Bankruptcy Court, substantially all of the assets of Spike Broadband Systems, Inc. of Nashua, NH for cash consideration of $4 million. Spike had been engaged in the development, manufacture, and sale of broadband wireless communications products and services until it filed for voluntary reorganization under Chapter 11 of the United States Bankruptcy Code in November 2001. MRF


DAVIS

CenturyTel Appoints Davis As Investor Relations VP

TONY DAVIS has joined CenturyTel as vice president of investor relations. Prior to taking the position at CenturyTel, Davis was employed as the CFO at docHarbor, Inc., an Internet document solutions subsidiary of Anacomp, Inc.

TDK Semiconductor Corp.—CHRIS ZENTNER to director of information technology; formerly manager.

Sharp Electronics Corp.—TERESA (TERRI) SIEBERT to associate director of marketing for Sharp's Appliance Division; formerly senior marketing manager.

Eleven Engineering—DR. CHARLES (CHANGSOO) KIM to vice president for Semiconductors; formerly president of the LG Electronics Institute of Technology in Seoul, Korea.

Adhesives Research, Inc.—SALLY D. GROOME to Midwest regional manager for the Industrial business unit; formerly employed in marketing and communications with the United Way in York County, PA.

CTS Corp.—RICHARD G. (DICK) CUTTER III to secretary; formerly vice president and assistant secretary. Also, JAMES L. CUMMINS to senior vice president for administration; formerly vice president of human resources. In addition, H. TYLER BUCHANAN to senior vice president; formerly vice president.

The Brattle Group—WILLIAM LINDSAY to senior advisor; formerly director of the Office of Electric Power Regulation at the Federal Energy Regulatory Commission (FERC).

Streaming21, Inc.—DAVID SILVER to president and CEO; formerly president and CEO of Kofax.

Touch America—PERRY COLE to senior vice president of revenue; formerly senior vice president of sales. Also, STEVE MAEDERER to vice president of sales; formerly vice president of broadband and wholesale sales.

Kivera—THOMAS V. BONOMI JR. to the position of CFO; formerly vice president of finance.

TÜV America, Inc.—CHRISTOPH WEIMER to CFO; formerly director of planning and analysis at Siemens Business Services, Inc.

Qualcomm, Inc.—JIM DOH to vice president and general manager of Qualcomm CDMA Technologies (QCT) Korea/Taiwan; formerly vice president of sales for Korea.

LCC International, Inc.—MARC WEINBERG to vice president of sales for the North America region; formerly headed the Eastern Region's sales team at Wireless Facilities, Inc.

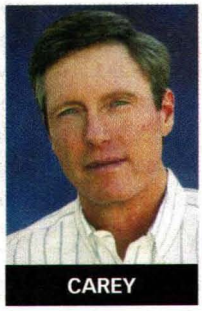
Current Analysis—KATE GERWIG to principal analyst for Telecom Services; formerly executive editor of *Tele.Com Magazine*.

Microchip Technology, Inc.—GANESH MOORTHY to the position of vice president of the Advanced Microcontroller and Automotive Division; formerly chairman and CEO of Cybercilium, Inc.

Yokogawa Corp. of America—SHUZO KAIHORI to CEO; formerly COO.

Racal Instruments, Inc.—TIM OSTROSKY to director of sales; formerly national sales manager for Integration.


OSTROSKY


CAREY

Celerity Systems, Inc.—TIM CAREY to vice president of marketing; formerly director of marketing for Anritsu Co., USA. MRF

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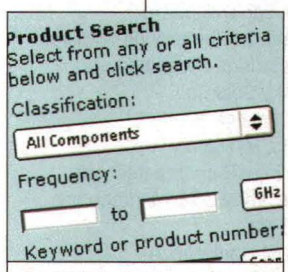

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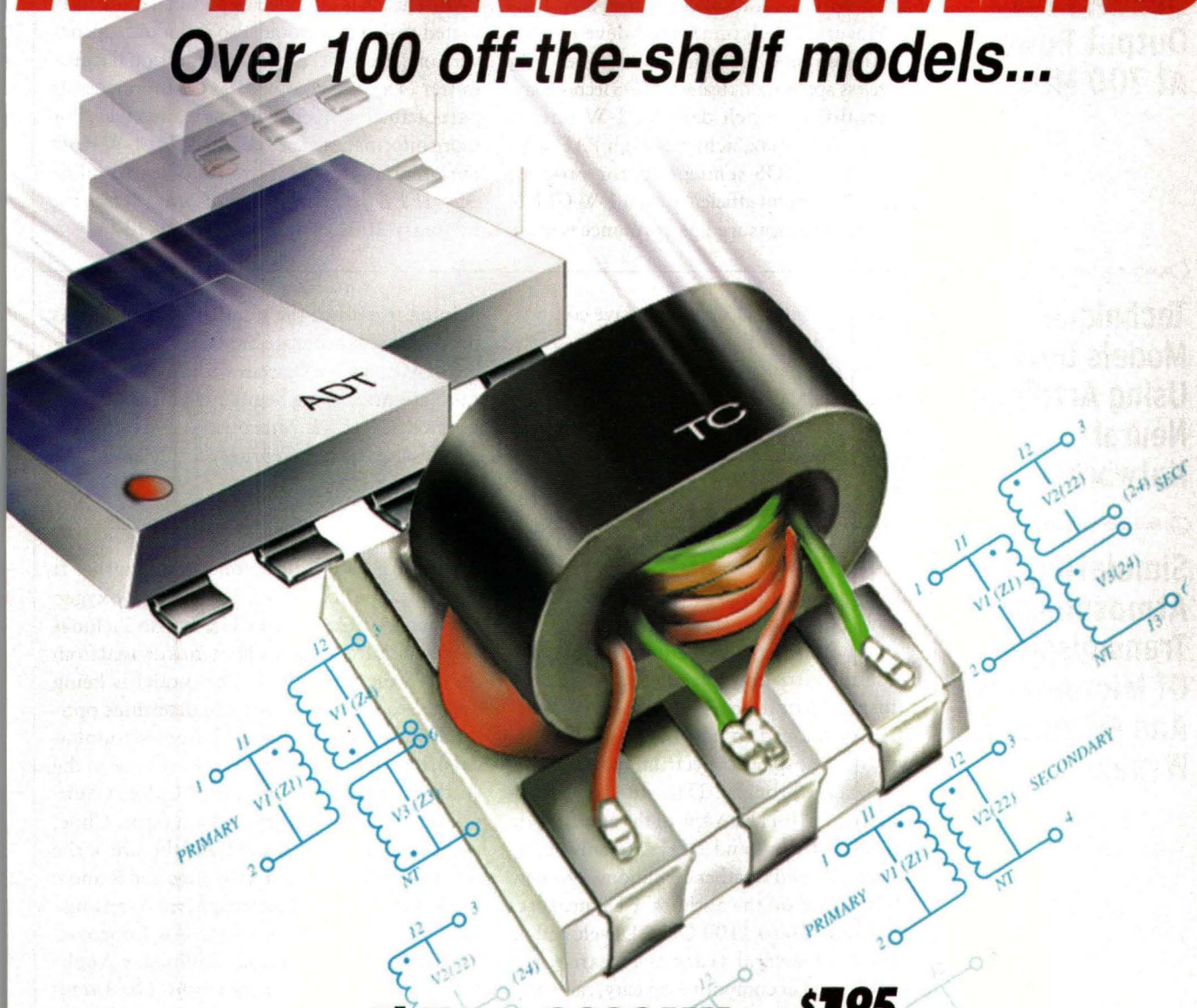

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CMOS PROCESS TECHNOLOGY is normally not associated with large-signal, high-frequency applications. But Koen Mertens and Michiel Steyaert of ESAT-MICAS, Katholieke Universiteit Leuven (Heverlee, Belgium) have developed a fully differential Class E power amplifier (PA) for wireless applications using CMOS technology. The amplifier, which delivers 1-W output power at 700 MHz, achieves high PAE with the silicon CMOS semiconductor process, reaching 62-percent efficiency for a 1-W GMSK spectrum. The measured performance is quite close to the authors' computer models for the CMOS amplifier, with simulated performance levels of 1.04-W output power and 66-percent PAE. The PA is intended as part of a more integrated design that would also include an upconversion (transmit) mixer and additional transmitter (Tx) and receiver (Rx) components as part of an all-CMOS wireless transceiver. For more information, see "A 700-MHz 1-W Fully Differential CMOS Class-E Power Amplifier," *IEEE Journal of Solid-State Circuits,* February 2002, Vol. 37, No. 2, p. 137.

Technique Models Devices Using Artificial Neural Networks

MICROWAVE CAE REQUIRES extensive computing power to analyze the EM response of a device. By using segmentation and finite-element methods, Jean Cid, Jesus Garcia, and Juan Zapata of the Universidad Politecnica de Madrid (Madrid, Spain) were able to develop a neural model for an arbitrary device, helping to reduce the number of data points needed for an accurate analysis. See "Modeling of Microwave Devices with Artificial Neural Networks Using Segmentation and Finite Elements," *IEEE Microwave And Optical Technology Letters,* February 5, 2002, Vol. 32, No. 3, p. 221.

Simulating The Atmospheric Transmission Of Microwaves And Millimeter Waves

THE ATMOSPHERIC TRANSMISSION of microwave energy is modeled to understand the propagation windows that support communications, radiotelescopy, and other applications. Juan Pardo, Jose Cernicharo, and Eugene Serabyn of the California Institute of Technology (Pasadena, CA) developed an improved model of the longwave atmospheric spectrum that is usable at frequencies through 2 THz, with some spectral lines up to 10 THz. A few of the atmospheric windows between 1 and 2 THz are only accessible under certain weather conditions. The new model is based on the authors' FTS measurements from 170 to 1100 GHz. Developed as the result of several years of research, the model is aimed at computing opacity, radiance, phase delay, and polarization along a particular path in the terrestrial atmosphere at millimeter and submillimeter wavelengths. It includes all effects due to water and oxygen resonances through 10 THz. It also includes semiempirical continuum-like terms derived from the FTS measurements. The model is being applied to analysis in order to determine optimum sites for new ground-based astronomical observatories. Three sites were used in the studies: Mauna, Hawaii (the Caltech Submillimeter Observatory), Chajnantor, Chile, and the geographic South Pole (the site of the Antarctic Submillimeter Telescope and Remote Observatory). See "Atmospheric Transmission at Microwaves (ATM): An Improved Model for Millimeter/Submillimeter Applications," *The IEEE Transactions On Antennas and Propagation,* December 2001, Vol. 49, No. 12, p. 1683.

InP HEMT MMIC LNA Bridges 65 To 110 GHz

DEMANDS FOR INCREASED BANDWIDTH, high-speed data transfers and near-instant Internet access have dramatically increased the frequency bands of interest for practical communications systems. In support of this, researchers at NASA's Jet Propulsion Laboratory (Pasadena, CA) have developed an InP HEMT low-noise amplifier (LNA) that is capable of operation from 65 to 110 GHz, and usable in both cryogenic and room-temperature systems. The amplifier incorporates devices from TRW (El Segundo, CA) with coplanar waveguide transmission lines, thin-film resistors, and thin-film capacitors mounted in a waveguide module. When operated cryogenically (at 24 K), a total of 17 of the LNAs yielded noise-temperature performance from 30 to 107 K at frequencies from 85 to 115 GHz. The corresponding noise temperatures at room-temperature noise temperatures were 250 to 470 K. For more information, see "InP HEMT MMIC Low-Noise Amplifier for 65 to 110 GHz," *NASA Tech Briefs,* February 2002, Vol. 26, No. 2, p. 37.

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2	0	SBB	5	800-2300	22-24	0.5-0.6	3.0-4.0	4.95
2	90	QBA	7	340-2400	21-28	0.25-0.80	3.0-7.0	6.95
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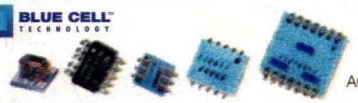

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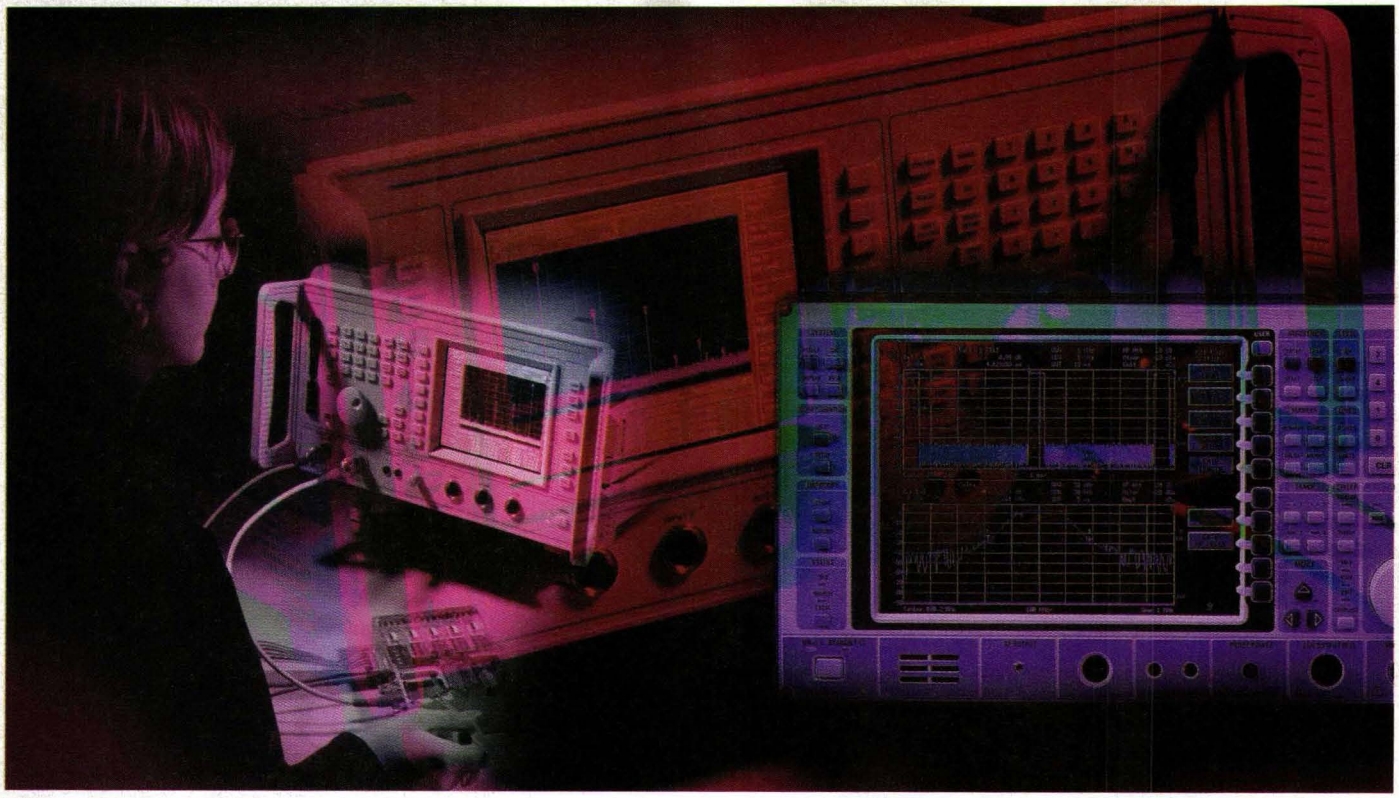

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DESIGN

Tracking Phase Noise In Short-Range Radios

Part 4 of this series on short-range radio will show how the signal source design plays an important role in the ultimate performance of a short-range radio system.

FARRON L. DACUS
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(480) 792-7017, e-mail: farron.dacus@microchip.com

JAN VAN NIEKERK
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Oscillator phase noise is often associated with receiver (Rx) sensitivity analyses. But in the design of a short-range radio system, especially those employing integrated low Q voltage-controlled-oscillator (VCO) methods, excessive phase noise can limit the effectiveness of a modulation scheme to transmit information. Too much phase noise can raise the effective bit-error rate (BER) of the system, and prevent transmitted information from being accurately received.

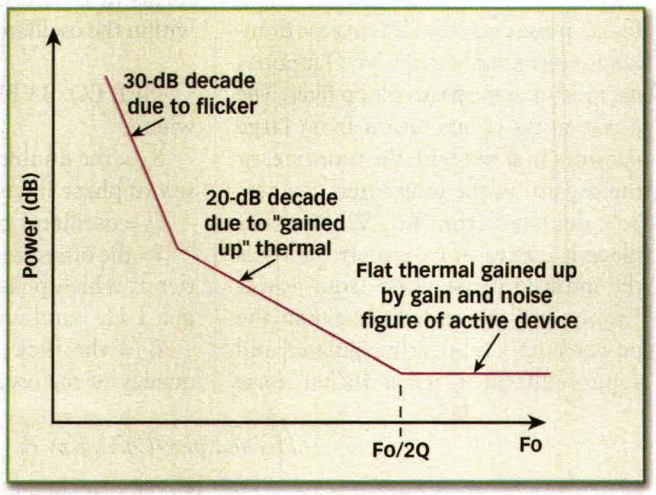
Oscillator phase noise may be intuitively defined as the noise attached to and spread around the carrier that is measured in the frequency domain. In this sense, it is similar to intended modulation and, if bad enough, interferes with desired modulation. It is generally measured on a per Hertz of frequency basis at some offset from the carrier and expressed in decibels relative to the total carrier power. Examination of the most important phase-noise issues of low-power VCOs is as follows. The induced "sideband-to-carrier" ratio, a peak-voltage magnitude-based ratio, is developed from small-signal frequency-modulation (FM) theory and is provided by:

[SEE EQ. 31 BELOW]

where:

$V_p(f)$ = the peak value of a sinusoidal baseband modulating voltage at frequency "f" on the steering input to the VCO and:

$$(Sideband\ mag/Carrier\ mag)(f) = \left[V_p(f)K_0\right]/2f \qquad (31)$$

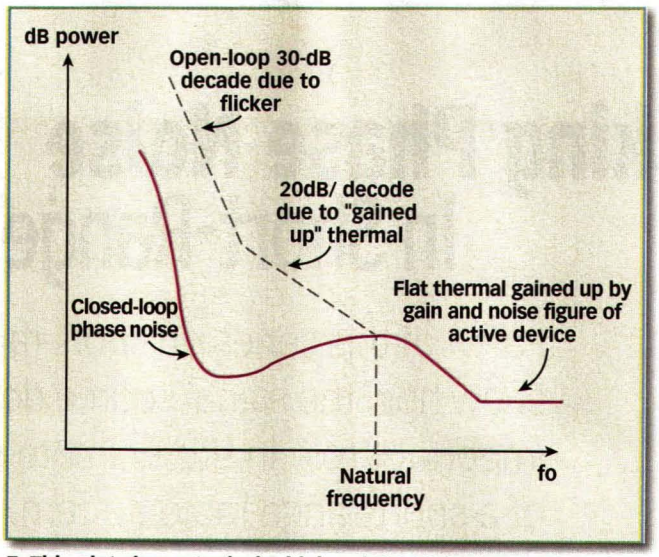


6. This plot shows basic VCO phase-noise characteristics.

K_0 = the gain of the VCO (in Hz/V.

The induced sidebands occur on both sides of the carrier. This equation is typically used to predict discrete spurs such as synthesizer sidebands (using Fourier series to obtain sinusoids), but it is also useful for broadband noise sources. A small transformation is needed to apply to these sources. Since $V_p(f)$ is a peak value and noise sources would typically be expressed in root mean square (RMS), it is possible to write the sideband to carrier power ratio based on RMS noise voltage, $V_n(f)$, as:

[SEE EQ. 32 BELOW]



7. This plot shows typical wideband PLL VCO phase noise. The integral of the area between the open- and closed-loop phase noise is the noise power removed by the PLL.

The sideband-to-carrier power ratio of Eq. 32 is the contribution of tuning input noise to the general phase noise L(f) that will be defined shortly. Taking 10log[S/C] will provide the contribution to phase noise from this source in units of decibels above the carrier per Hertz. An example of the effect of Eq. 32 is the phase-noise effect of the flicker noise of an operational amplifier that is used in a loop filter or buffer driving a VCO. This seldom-mentioned unpleasant fact is actually one of the primary reasons behind the modern prevalence of the current-pump PLL. The current pump is off more than 99 percent of the time, greatly reducing its flicker noise and often leaving the dominant noise source in the VCO input as the resistor in the passive loop filter. The phase-noise contribution from large resistors in series with the tune line, or the resistor in the loop filter, can also be calculated from Eq. 32. Another place it comes in extremely handy is the induced phase noise from power supply and ground. Even when the power supply is linearly regulated and supposedly quiet, it has flicker noise that can dominate the phase-noise profile. To calculate this noise, one uses Eq. 32 where K_0 becomes K_{0p}, the frequency change per volt of power-supply change. This term is usually only a decade or so down from K_0, and can easily dominate. Many a designer has done a good job on the standard phase-noise control issues and has then been unpleasantly surprised by a power-noise-dominated phase noise up to several tens of decibels worse than predicted.

The oscillator design itself is inherently phase noisy even when the tune and power lines are sufficiently quiet. A relatively accurate expression for oscillator phase noise (derived from forms in refs. 6 and 7) from factors within the oscillator is given by Eq. 33:

[SEE EQ. 33 BELOW]

where:

S_ϕ = the double-sided spectral density of phase fluctuation (in rad^2/Hz),

f_0 = oscillator operating frequency,

f = the offset frequency from the carrier at which phase noise is measured in a 1-Hz bandwidth,

f_C = the flicker-noise corner frequency of the oscillator active device,

the frequency where flicker noise (1/f noise) on the output of the device is equal to the thermal floor multiplied by gain and noise factor,

Q = loaded Q of the resonator,

G = oscillator active-device gain in compression,

F = oscillator active-device noise factor (not in decibels) in compression, typically higher than the uncompressed noise factor,

kT = Boltzman's constant and absolute temperature, and

P_0 = the output power of the oscillator's active device.

The term S_ϕ shall be momentarily related to what is commonly referred to as phase noise. This equation is basically Leeson's phase-noise model with a flicker-noise corner added, and with power on the output side of the amplifier to more clearly show the effect of active-device gain G.

The somewhat nonintuitive term "spectral density of phase fluctuation" may be, using a small angle approximation, interpreted as the noise spectral-density ratio relative to carrier power commonly known as phase noise. Phase noise is sometimes provided as $L(F_0) = 10\log(S_\phi/2)$ in units of decibels above the carrier per Hertz. Since linear and decibel units can be considered here, it might be wise to make more clear definitions:

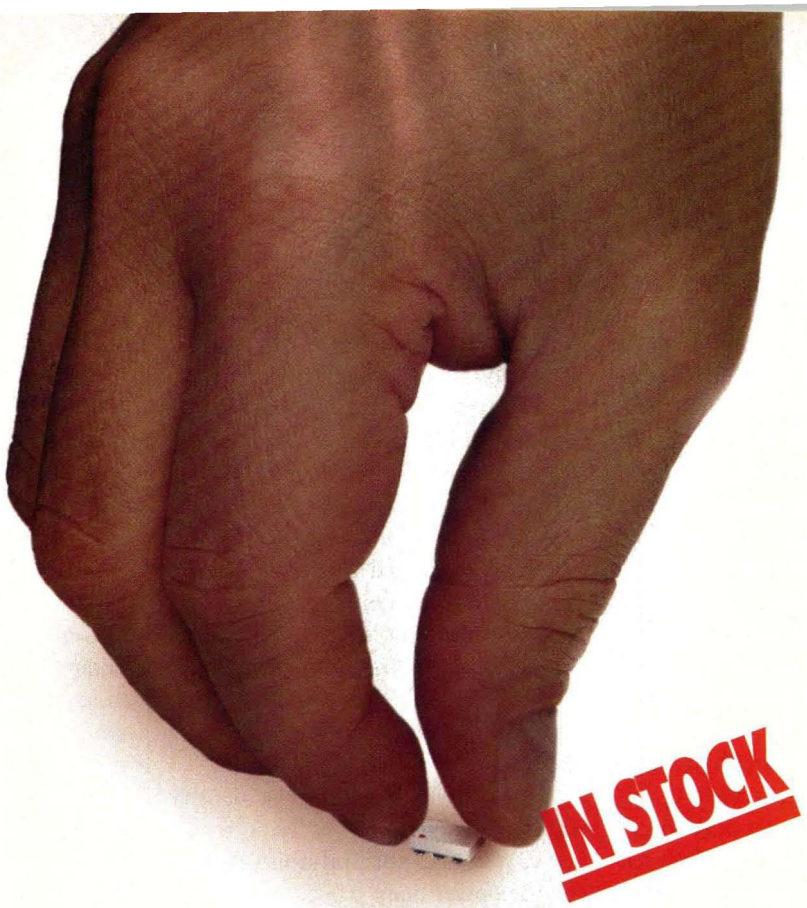
$$L(f) = S_\phi(f)/2 \qquad (34)$$

$$L(dB)(f) = 10\log\left[S_\phi(f)/2\right] \qquad (35)$$

The factor of 1/2 comes from the convention that phase noise as a power-spectral density is real and observed on one side, while phase-fluctuation spectral density has double the power in that it represents the phase fluctuation in a 1-Hz bandwidth on both sides of the carrier.

$$(Sideband\ pwr/Carrier\ pwr)(f) = \left\{\left[V_n(f)K_0\right]/\left(\sqrt{2}f\right)\right\}^2 \qquad (32)$$

$$S_\phi(f) = \left[(f_o/2Q)^2(GFkT/P_0)f_c/f^3\right] + \left[(f_o/2Q)^2(GFkT/P_0)/f^2\right] + \left[(GFkT/P_0)f_c/f\right] + \left[(GFkT/P_0)\right] \qquad (33)$$



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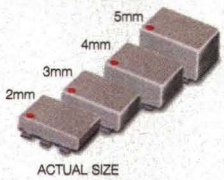

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ADE-12	+7	50-1000	7.0	35	17	2	2.95
ADE-4	+7	200-1000	6.8	53	15	3	4.25
ADE-14	+7	800-1000	7.4	32	17	2	3.25
ADE-901	+7	800-1000	5.9	32	13	3	2.95
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ADE-5X	+7	5-1500	6.2	33	8	3	2.95
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ADE-11X	+7	10-2000	7.1	36	9	3	1.99▲
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ADE-28	+7	1500-2800	5.1	30	8	3	5.95
ADE-30	+7	200-3000	4.5	35	14	3	6.95
ADE-32	+7	2500-3200	5.4	29	15	3	6.95
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$$L(f) = \left[(f_o/2Q)^2 (GFkT/2P_0) f_c/f^3\right] + \left\{(f_o/2Q)^2 (GFkT/2P_0) + \left[(V_{n1}(f)K_0)/\sqrt{2}\right]^2 + \left[(V_{n2}(f)K_{0p})/\sqrt{2}\right]^2\right\}/\left[f^2\right] + \left[(GFkT/2P_0) f_c/f\right] + (GFkT/2P_0) \quad (36)$$

A typical VCO free-running phase noise shape is shown in **Fig. 6,** where it will be noted there are only three main regions despite the fact that Eq. 33 has four terms. The 1/f term has been left out since the $1/f^2$ term dominates it in most cases. For bipolar transistors, the flicker corner can be from below 100 Hz to several kilohertz. For complementary metal-oxide semiconductor (CMOS), the flicker corner is typically from several tens of kilohertz to 1 MHz or slightly more. The $1/f^2$ rise in phase noise begins at the resonator half-bandwidth, or $f_0/2Q$. For the typical Q values and center frequencies of RF oscillators, this is well-beyond the 1/f corner, so the term 1/f never catches up. However, for low-frequency, high-Q oscillators such as crystal oscillators, the 1/f term can be noticed, particularly for CMOS with its high flicker noise.

Equation 33 may be extended to include the effects of the tuning and power-supply noise by adding the appropriate forms of Eq. 32 to Eq. 33. This is a requirement for the high-gain VCOs typical of integrated short-range radios. Doing this and converting to phase noise in watts (noise) per watt (carrier) per Hertz in Eq. 34 yields Eq. 36:

[SEE EQ. 36 ABOVE]

where:

V_{n1} = RMS spectral-noise density on the tune line,

V_{n2} = RMS spectral-noise density on the power line,

K_0 = VCO gain (in Hz/V), and

K_{0p} = the power-supply-pulling susceptibility of the VCO (in Hz/V).

This equation allows quick visualization of the design methods used to minimize oscillator phase-noise noise. These are to maximize loaded Q, maximize output power (simultaneous voltage and current compression in the active device will extract the most power from the available budget), minimize flicker factor (bipolar is superior to FETs), minimize loop gain (3 to 6 dB over loop losses), minimize compressed noise figure (minimum compression helps for this), minimize VCO gain and input noise, and minimize pulling frequency susceptibility and noise on the supply.

Parameter V_{n1} is primarily resistive thermal noise and active device flicker noise. If an opamp drives a VCO input, its flicker noise is likely to dominate

the phase noise, especially for low-power CMOS opamps. For current-pump-based PLLs, the flicker noise will be low since the pulse width of the current pumps will approach zero (unless the VCO is directly FSK modulated, when the phase detector will encode this modulation when the loop attempts to hold the VCO exactly on frequency). Thus, for current-pump PLLs, V_{n1} may be primarily thermal noise. Given the typically small current pump values, and typically wide loop bandwidth in short-range radios, this resistor is typically much larger than that used in the loop filter of a design such as a cellular telephone synthesizer. With the high VCO gain typical of short-range radios, its thermal noise is often quite noticeable. It is calculated using:

$$V_n(rms) = (4kTR)^{0.5} \qquad (37)$$

The common practice of following a second-order loop filter (two capacitors, one resistor, which makes a third-order loop) with a second RC stage (resulting in a fourth-order loop) must be viewed with caution in short-range radios. To avoid loading the loop filter with this last stage, this second resistor is normally made 5 to 10 times larger than the damping resistor and thus it has that much more thermal noise. Equations 36 and 37 in concert with the closed-loop phase-noise analysis described below should be used to check if this additional resistor and even the standard loop resistor are acceptable.

The earlier statement that the wideband PLL could cover noise problems in the VCO will now be proved and analyzed. From Eq. 32 it is clear that phase noise may be referred to the input as a noise voltage in a way analogous to how noise in amplifiers can be referred to their input. If the oscillator is imagined as noiseless and all phase noise is induced by an imaginary RMS VCO open-loop noise voltage, V_{nvo}, on the VCO tune input, and noting that total sideband to carrier power ratio is the same as L(f), then Eq. 32 may be solved for V_{nvo} as:

$$V_{nvo} = \left\{ f\left[2L(f)\right]^{0.5} / K_0 \right\} \qquad (38)$$

This is the input-referenced open-loop VCO noise, where L(f) is given by Eq. 36. However, the open-loop input-referenced noise will be modified by the closed-loop action of the PLL. The PLL will reduce the noise within the loop bandwidth in the attempt to keep the phase error equal to zero. It may be shown by basic analysis of Fig. 4 (in Part 3) that the effect of the loop on the open-loop input-referenced VCO phase noise is exactly equal to multiplying by He(s). That is, the injected noise V_{nvo} that models free-running phase noise is modified by the loop to be the RMS closed-loop noise voltage quan-

tity V_{nvc} residing directly on the tune voltage, and this voltage is found by:

$$V_{nvc} = V_{nvo}|H_e(s)| \qquad (39)$$

This function can be used in the standard second-order normalized form given earlier, or for high accuracy in higher-order loops, it can be calculated based on the full set of intended and parasitic poles in the loop.

Next, the question of crystal-oscillator phase noise and its effect will be addressed. Although a crystal oscillator exhibits high Q and inherently low phase noise, its phase noise is much worse in CMOS high flicker processes, and is it then multiplied by divider value N through the closed-loop action of the PLL. While the crystal reference for a cellular telephone is typically a well-optimized bipolar device consuming about 2 mA, the crystal oscillator for a short-range transmitter typically consumes 200 µA and is often implemented with a CMOS digital gate active device that has typically higher noise figure and much higher flicker noise than a bipolar transistor. Referring to Eq. 36, the increase in phase noise for this short-range reference (compared to a cellular telephone reference) would typically be on the order of 20 to 40 dB. This combination of factors is such that effect of the phase noise of the crystal oscillator on the total closed-loop phase noise is definitely not negligible for short-range FSK systems. Analysis of the transfer functions of the closed-loop PLL will show that:

$$V_{nxc} = V_{nx}\left(K_{ox}N/MK_0\right)|H(s)| \qquad (40)$$

where:

V_{nxc} = the closed-loop RMS noise that appears on the VCO input (not the VCXO input) as a result of crystal-oscillator phase noise,

V_{nx} = crystal-oscillator noise referred to the crystal-oscillator steering input (a VCXO),

K_{ox} = the crystal-oscillator tune slope (in Hz/V),

M = the value of any divider between the crystal oscillator and the phase detector, and

H(s) = the phase-transfer function.

The method of representing crystal-oscillator phase noise referred to a tune input is useful because many references are VCXOs to allow exact frequency trim, which is then susceptible to noise on the tune line from thermal, flicker, and power-supply noise sources that should be taken into account. The effect of supply noise on the crystal oscillator is taken into account just as it was for the VCO, though K_0 and K_{0p} for the crystal will be much lower than the corresponding tune slopes for the VCO. If the crystal oscillator is not a VCXO, then

an arbitrary value for K_{ox} can be assumed for the purpose of this calculation.

The total noise from the VCO, V_{ntc}, and the crystal oscillator referred to the input of the VCO from these rms referred to VCO input voltages in the closed-loop state is:

$$V_{ntc} = \left(V_{nvc}^2 + V_{nxc}^2\right)^{0.5} \quad (41)$$

To get the total resulting closed-loop phase noise, V_{ntc} is then applied back through Eq. 32 to get

$$Lc(f) = \left\{\left[V_{ntc}(f)K_0\right]/\sqrt{2}f\right\}^2 \quad (42)$$

The effect of the closed loop on PLL noise is especially dramatic for a wideband loop **(Fig. 7)**. Well inside the loop's natural frequency, the noise is suppressed 40 dB per decade (a second-order magnitude transfer function). If the loop natural frequency is out where the VCO phase noise is at 20 dB per decade, and other noise sources do not limit loop action, then the phase noise will decline 20 dB per decade over that frequency segment. A common occurrence is for divider noise (not shown in this model) to limit this effect, typically so that the phase noise approximately flattens out at some offset well inside the loop bandwidth. Also, the loop can only reduce noise to the point of the multiplied crystal reference noise as given by Eq. 40, which is why the phase noise turns around and starts rising again. However, crystal oscillators have very low phase noise due to their very high Q, and so despite degradations such as divider noise a wideband PLL can provide a high degree of phase-noise clean up. It is this action that allows low Q integrated VCOs such as relaxation oscillators, usually running on unregulated supplies, to provide adequate performance for FSK modulation.

As an example, the model presented above has been implemented as a MathCAD model to analyze the performance of a typical BiCMOS transmitter implementation. The following basic parameters apply. The crystal oscillator has a frequency of 13.5625 MHz, a loaded Q of 5000, a flicker corner frequency of 1 kHz, an active device noise factor of 10, gain of 6 dB, current consumption of 250 µA, Z_{out} of 2000 Ω, and output power of 63 µW. The VCO has a frequency of 434 MHz, a Q of 1 (a relaxation oscillator), a flicker corner frequency of 1 kHz, an active device noise factor of 10, compressed gain of 6 dB, current consumption of 3 mA, output impedance of 300 Ω, output power of 1.35 mW, and K_0 = 200 MHz/V. The PLL has a natural frequency of 50 kHz, a damping factor of 1.2, current-pump cur-

rent of 260 μA, N = 32, M = 1, filter R = 470 Ω, and filter C = 0.015 μF.

In the long version of this article on www.mwrf.com, phase-noise plots are also shown for a crystal oscillator, an open-loop VCO, and a closed-loop VCO for the situation that was just described. For the crystal oscillator with its very high Q, the $1/f^3$ term is the first to rise above the floor. The VCO phase-noise curve rising above 0 dB at very low offset frequency is not what actually happens, as physically L(dB)(f) must flatten at 0 dB. This error is due to the 1-Hz granularity assumed in the phase noise not being adequate as the offset approaches 1 Hz, which is particularly apparent in this low-Q, high-flicker-noise case. The closed-loop noise model shows the effect of the 50-kHz-wide loop on the poor phase noise of the relaxation VCO. The phase noise at offsets lower than 1 kHz is equal to the crystal-oscillator phase noise plus $10\log N^2$ dB, as would be expected from the PLL multiplying the reference by N (here N = 32). This phase noise will limit demodulated FSK signal to noise ratio to approximately 28 dB no matter how strong the receiver input signal to noise power ratio (SNR).

Short-range radio systems are almost totally ASK or FSK modulation, since these modes are the lowest power and simplest to implement. For ASK systems phase noise is not a serious issue unless it is so bad that it places noticeable energy outside the receive bandwidth or fails regulatory requirements (see Part 2 of this article series, October 2001, p. 79). But for digital FSK (becoming especially popular in Europe) the phase-noise sets an upper limit on the signal-to-noise ratio that may be achieved. It is desired for this limit to be high enough that in practice it is not a noticeable factor in BER, which generally calls for the this limit to be 20 dB or more.

If a phase noisy unmodulated carrier is detected with a sensitive FM demodulator, the output will display a noise referred to as the residual frequency modulation, which is a noise that competes with the desired FSK. For integrated phase noise less than 1 rad^2, the square of RMS residual FM due to phase noise over the bandwidth f_a to f_b is given by (ref. 7):

$$\Delta f^2 = 2\int_{f_a}^{f_b} f^2 L(f)df \qquad (43)$$

This noise sets a limit on FSK signal-to-noise ratio (SNR) for intended FSK expressed as one-sided RMS frequency deviation f_{rms} that is given by:

[SEE EQ. 44 BELOW]

The limits of integration selected in Eq. 43 depend on the data rate and protocol used in the system, and the acceptable BER. The cascade of baseband filtering in the transmitter and receiver generally sets these limits. Sometimes this filtering may be pure lowpass and extend all the way to DC, but it is common for something in the system or circuit design to force a low-frequency highpass function such that the baseband filtering is actually bandpass. For example, FSK PLL modulation imposed by direct modulation of the VCO with correcting integrator to reduce distortion (see US patent No. 6172579) cannot modulate all the way to DC. The rfPIC12C509 with FSK via the crystal reference does go all the way to DC, but the receiver may not necessarily go all the way to DC. For example, the Rx demodulator may be highpass filtered to remove DC offsets. Rx automatic frequency control (AFC) also sets a lower limit on the frequency content of the demodulated FSK output, the effect of which is to place one or more highpass poles at the AFC system bandwidth. If any of these highpass poles are present, they suppress phase noise below the poles and provide a good number to use for f_a in Eq. 43. However, if the protocol used has noticeable low frequency content, then AC coupling in the system above this content will degrade BER by removing desired energy. For example, at 20 kb/s using a non-return-to-zero (NRZ) protocol an AC-coupled highpass response will degrade BER the equivalent of only about 1/10 of a dB of SNR for a 10-Hz corner frequency, but at 50 Hz will degrade BER by about 1-dB equivalent reduction in SNR. A Manchester format with zero DC content would have less susceptibility to high pass coupling in the system. A rule of thumb for selecting f_a for DC-coupled systems is to set it a 0.1 percent for NRZ and 1 percent for Manchester, for which accuracy should be to within a small fraction of 1 dB. For the upper limit of integration f_b, select the lowest pole of the baseband lowpass filtering, typically about one-half the data rate for binary FSK.

The phase-noise-limited FSK signal to noise ratio may now be examined, with highly enlightening results for the design of integrated short-range transmitters. Incorporating Eqs. 43 and 44 into a MathCad model allows running up a variety of cases of process flicker noise, VCO Q, and the PLL parameters. The final result of this is to give the necessary PLL bandwidth to provide a minimum acceptable SNR. The case examined is a data rate of 20 kb/s, NRZ formatted, with a frequency deviation of 20 kHz peak-to-peak (modulation index = peak-to-peak deviation/data rate = 1.0), with lowpass filtering at 10 kHz (bandwidth-time product BT = 0.5). The results are shown in detail in a table in the long-form web version of this article. Key results include the facts that with Q = 1, integrated VCOs attaining sufficiently low phase noise for acceptable FSK require that the loop natural frequency be at least 30 kHz for typical BiCMOS processes. For typical CMOS processes, the loop natural frequency should be at least 65 kHz.

Next month, Part 5 of this article series will detail modeling methods for loop antennas, the idiosyncrasies of modulated crystal oscillators, and practical issues of designing short-range radios to meet Federal Communications Commission (FCC) regulations. MRF

$$(S/N)[phase\ noise\ bounded] = 10\log\left(f_{rms}^2/\Delta f^2\right) \qquad (44)$$

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DESIGN

Linearity/Power Considerations In 5-GHz WLANs

Achieving an optimum combination of linearity and high power in WLAN systems requires design trade-offs that could benefit from a combination of methods.

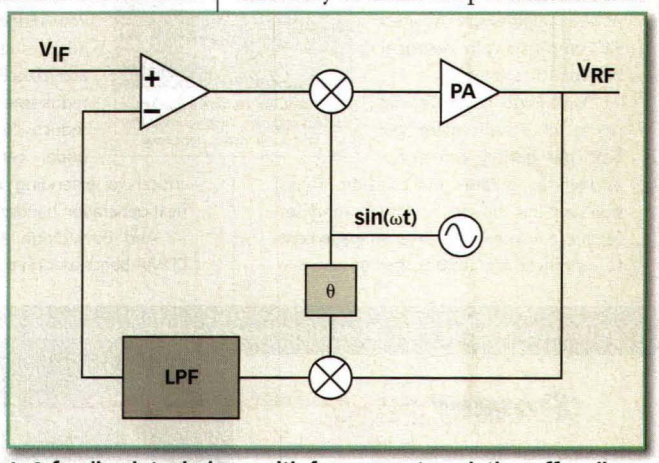
J.S. WIGHT
Principal Architect
IceFyre Semiconductor, 300-411 Legget Dr., Kanata, Ontario K2K 3C9, Canada; (613) 599-3000, Internet: www.IceFyre.com.

Multicarrier modulation schemes such as orthogonal frequency division multiplexing (OFDM) impose stringent linearity requirements on the RF portions of a transceiver. For the transmitter (Tx)—specifically the power amplifier (PA)—this strong linearity requirement is further complicated due to the additional need for high power levels to meet wireless local-area-network (WLAN) range requirements.

Many communications devices are designed to deliver linear output to ensure peak performance, but all active devices go through compression at certain power levels, causing some signal distortion. Strong linearity is important for OFDM schemes because there is a high peak-to-average power ratio. Therefore, for the systems designer, there is a trade-off to be made between linearity and power.

Several approaches are available to overcome these difficulties in RF PAs, including feedback, feedforward, polar decomposition, predistortion, and linear amplification with nonlinear components (LINC) techniques. However, due to the wide bandwidths of IEEE standard 802.11a and HyperLAN-2 WLAN systems, each of these approaches is hampered by some limitations. For instance, feedback techniques can exhibit stability problems, feedforward and polar decomposition techniques can have matching issues, predistortion can have dynamic range limitations, and LINC can exhibit significant loss. To serve the new 5-GHz WLANs coming down the line, it is necessary to employ a new approach based on a combination of techniques.

Feedback techniques are not easy to incorporate in RF PAs due to their tendency to cause loop instabilities. In



1. A feedback technique with frequency translation offers limited advantages for a 5-GHz WLAN PA.

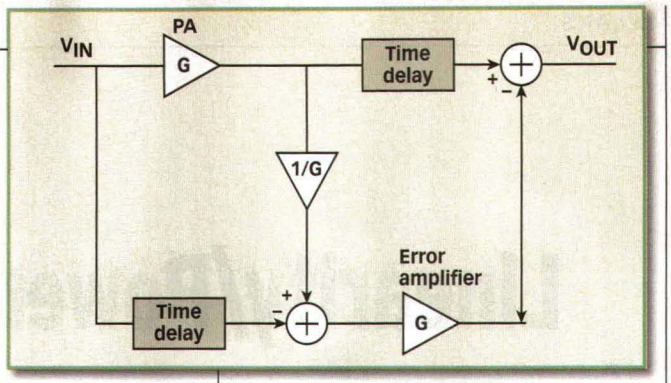
addition, for a highly nonlinear PA, a high loop gain must be realized to provide the linearization required. The major design challenge associated with this approach is overcoming loop instabilities (arising from various resonances of parasitic couplings, package parasitics, and transient currents) which make feedback PAs subject to spontaneous oscillation.

Feedback techniques can, however, be applied successfully to a combined upconverter/PA. The problem of high loop gain is alleviated by allocating most of the loop gain to the lower intermediate frequency (IF), where resonance is less prevalent. This is shown in **Fig. 1**, where a portion of the output power is downconverted to the original IF. In effect, the loop attempts to make the modulation of V_{RF} a replica of the modulation of V_{IF}.

To ensure stability, however, the closed-loop phase must not approach 180 deg. for any frequency having a loop gain greater than 1. To achieve this, the phase quadrature (θ) of the reference local-oscillator (LO) frequency used to downconvert the RF is set to ensure a safe margin. A significant problem with this approach is the dependence of θ on temperature, processing parameters, and output power, which makes it difficult to guarantee stability.

In the past, designers also tried to use feedback techniques for Txs by incorporating an independent upconversion of quadrature signals that is then applied to a common PA. In this case, the quadrature downconversion in the feedback loop converts the RF back into two quadrature IF components, ready for comparison with the two original quadrature IF signals. Due to the added complexity and sensitivity to temperature and process, this architecture is not commonly used. As a result of the difficult stability issues, the use of feedback for improving the linearity of PAs for 5-GHz WLAN applications is not considered to be very practical.

Figure 2 shows a typical feedforward approach for improving linearity in a



2. The two amplifiers in feedforward architectures exhibit substantial phase shifts at high frequencies, which must be compensated for with two true-time delay elements.

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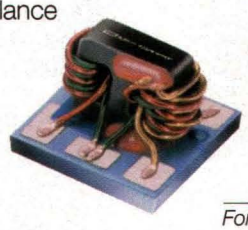

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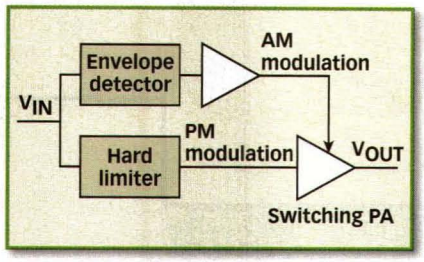
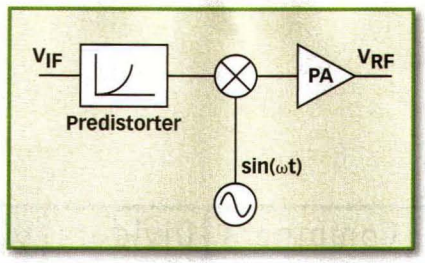


3. In a polar-decomposition technique, the input signal is applied to an envelope detector (to generate the envelope-modulated signal) and a hard limiter (to generate the PM signal).



4. Predistortion techniques require initial and ongoing calibration to select the appropriate nonlinearities for the magnitude and phase of the signal.

PA. In this technique, the output voltage waveform from a nonlinear PA is the summation of an amplified replica of the input signal and an error signal. The feedforward architecture determines this error and subtracts it from the amplified output waveform. An attenuator having a factor equal to the gain of the PA attenuates a portion of the output signal. This attenuated output signal is compared with the original input signal to create an error signal. Finally, this error signal is amplified by the same gain as the PA, and subtracted from the PA's output signal.

Unfortunately, at high frequencies—such as 5 GHz—the two amplifiers in the feedforward architecture exhibit substantial phase shifts. This can be compensated for by the use of two true-time delay elements as shown in Fig. 2.

Unlike feedback architectures, feedforward techniques are inherently stable, even with substantial phase shifts in each component. However, the passive true-time delay elements are lossy, and the degree of linearity achieved is dependent on the gain and phase (true-time delay) matching of the signals applied to each subtractor. For example, with a gain mismatch of 5 percent and a phase mismatch of 5 deg., the suppression of the power in the intermodulation (IM) products will be limited to only 20 dB. To add to the challenge, to prevent reduction of the overall output power, the output subtractor must have low power loss.

Due to the difficulty in achieving the substantial tracking accuracy required, as well as the inherent challenges in realizing low loss, true-time delay elements and combiners, feedforward techniques are not easily implemented for 5-GHz WLANs.

Any bandpass signal modulation can be represented as simultaneous and independent amplitude modulation (AM) and phase modulation (PM). As a result, a signal can be decomposed into an envelope-modulated signal and a PM signal. These two modulated signals can be amplified separately and later combined. This is known as polar decomposition.

In this method **(Fig. 3)**, the input signal is applied to an envelope detector (to generate the envelope-modulated signal) and a hard limiter (to generate the PM signal). The decomposed signals are individually amplified and combined in the PA. Since the PM signal is constant envelope, it can be amplified with a switching-mode (Class D or E) PA that features very-high efficiency. In addition, the envelope-modulated signal can control the amplitude of the switching-mode PA through adjustment of its bias.

Compared with feedback and feedforward approaches, polar decomposition has an advantage since it does not require linear PAs. However, any mismatch between the phases and gains of the two decomposed signal paths must be minimized. This is difficult since the two paths consist of vastly different types of circuit elements, working at widely different frequencies.

A further complication results from AM-to-PM conversion at high frequencies. In the PM signal path, the limiters will pass some residual AM. If

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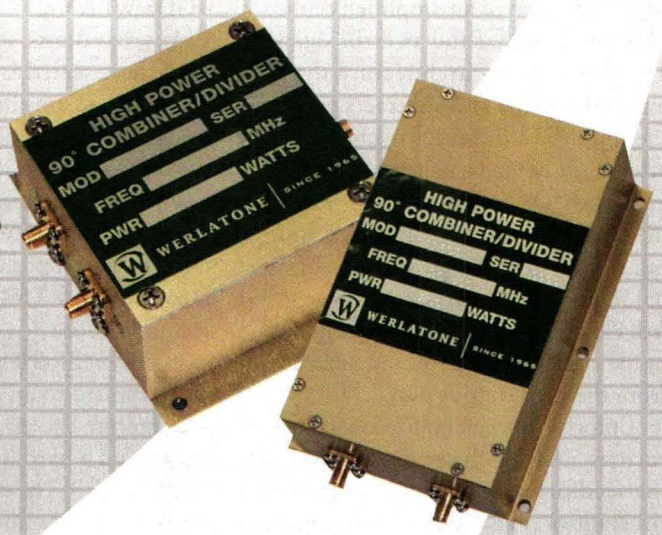

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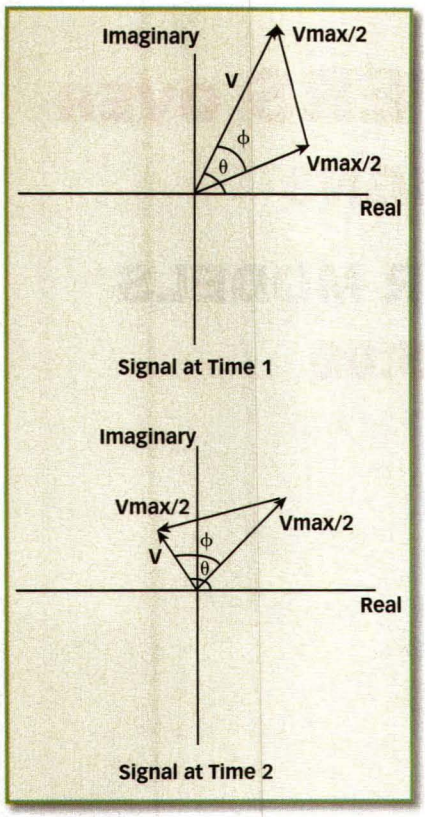


5. LINC decomposes a single bandpass signal into two constant-amplitude PM signals.

this residual AM passes through subsequent circuits that have unsymmetrical insertion phase (with respect to the carrier frequency), it will be transformed into PM sidebands. Similarly, in the AM signal path, a variation of the drain-junction capacitance of the switching transistor with the value of the envelope-modulated signal, will result in a change in the phase of the signal produced by the switching transistor.

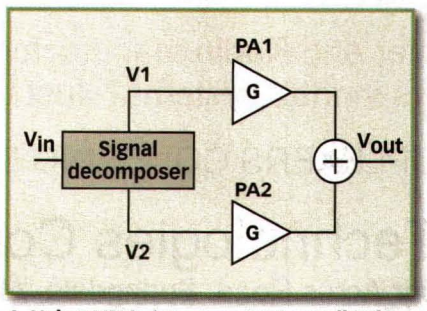
An additional concern arises in the loss of efficiency associated with the circuitry needed to control the bias voltage. The entire supply current of the PA will flow through the controlling transistor, which, as a result of the varying voltage across it, dissipates a considerable amount of power. Due to the rapid changes in amplitude of OFDM signals and the difficulty in accurately tracking them with an adjustable bias circuit, polar decomposition techniques are also not well-suited for use in 5-GHz WLANs.

Predistortion architectures are designed to provide the inverse input/output (I/O) function of the saturating PA. In effect, the predistortion circuit enables enhanced gain for large amplitude signals before they are applied to the PA. Further, the predistortion circuit attempts to compensate for any amplitude-dependent insertion phase by providing the opposite phase change.

These circuits can operate at the RF of the PA, or they can operate at an IF before upconversion, as shown in **Fig. 4**. When predistortion is applied at the RF, the correct nonlinearity for the magnitude of the I/O transfer function can be obtained with an appropriately biased pair of parallel, back-to-back Shottky diodes, while the necessary insertion-phase can be realized with a similar configuration of suitably biased varactor diodes. When predistortion is applied before upconversion, it can be realized as either an analog or digital implementation.

An advantage of this approach is that it does not incorporate any closed feedback loops and therefore does not have stability problems. However, achieving an accurate inverse I/O function of the PA is difficult. In fact, accurate predistortion requires initial and ongoing calibration to select the appropriate nonlinearities for the magnitude and the phase, which adds a significant degree of complexity to this approach. As a result, stand-alone predistortion techniques are not well-suited to the OFDM signals used in 5-GHz WLANs.

LINC is a relatively new technique



6. Using LINC, two constant-amplitude signals are amplified with nonlinear PAs, and subsequently combined to reform an amplified version of the original signal.

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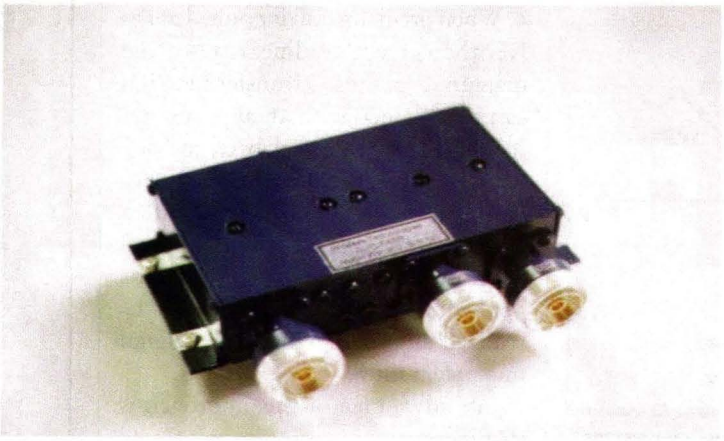

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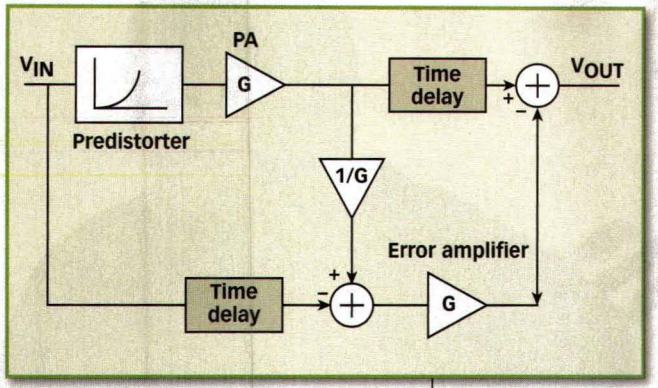

7. Inserting a predistortion element into a feedforward linearization architecture can help overcome the difficulties facing each technique individually.

that decomposes a bandpass signal into two constant-amplitude PM signals (**Fig. 5**). The two constant-amplitude signals are then amplified with nonlinear PAs, and subsequently combined to reform an amplified version of the original signal, as shown in **Fig. 6**.

The major design challenge when using LINC is determining the rate at which the phases for the two constant-amplitude signals must be updated. It is also very important to avoid gain and phase mismatch between the two signal paths, since this will result in signal distortion. Finally, the output combiner will introduce a significant loss if it is to achieve a high isolation between the PAs, so this must be accounted for in the design. Due to the inherent loss associated with LINC for signals having high peak-to-average power ratios, it is also not a good choice for improving the linearity in an OFDM system.

Due to the limitations of each of these traditional methods, it may be useful to combine certain aspects of the individual linearization techniques to provide a high degree of linearity in the analog portions of a multicarrier OFDM transceiver.

For example, inserting a predistortion element into a feedforward linearization architecture (**Fig. 7**) may help overcome the difficulties facing each technique individually. In particular, the predistortion element can reduce the magnitude of the error signal in the feedforward architecture, which will help alleviate its tracking-accuracy requirement. Similarly, the feedforward architecture could help reduce the accuracy required in the inverse I/O function of the predistortion element.

Similarly, a LINC architecture could be preceded with a polar-decomposition engine. Here, the polar-decomposition engine would be adapted to reduce, but not remove, the AM on its PM signal component. This signal component could then be applied to the LINC, with the residual AM signal component from the polar-decomposition engine being reintroduced as a bias adjustment of the LINC switching-mode PA. This combination technique addresses the difficulty of rapid changes in signal amplitude experienced by polar decomposition, and the inherent loss experienced by LINC, by partitioning the total amplitude variation of the signal into two components, and applying the separate components individually to the two architectures. There are other combinations beyond these examples that warrant exploration and may provide additional synergistic performance.

Improving linearity in PAs for radios using modulation schemes with large peak-to-average ratios is extremely important to achieve high efficiency and, hence, low power consumption. While there are several conventional methods that can be used for this, each technique has its own set of limitations and design challenges. Addressing these limitations is the key to the successful realization of any efficient PA. By moving beyond conventional approaches, it is useful to combine certain aspects of multiple techniques to produce variants well-suited for 5-GHz WLAN applications. MRF

FOR FURTHER READING

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2. T.H. Lee, *The Design of CMOS Radio-Frequency Integrated Circuits*, Cambridge University Press, Cambridge, England, 1998.

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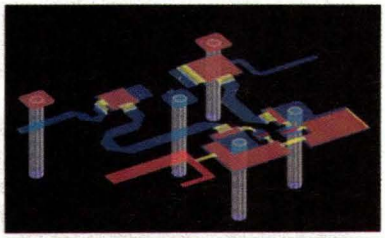

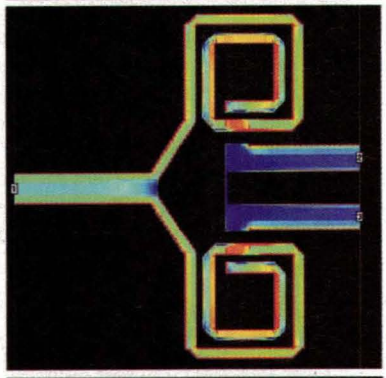

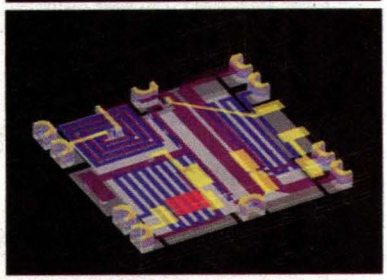

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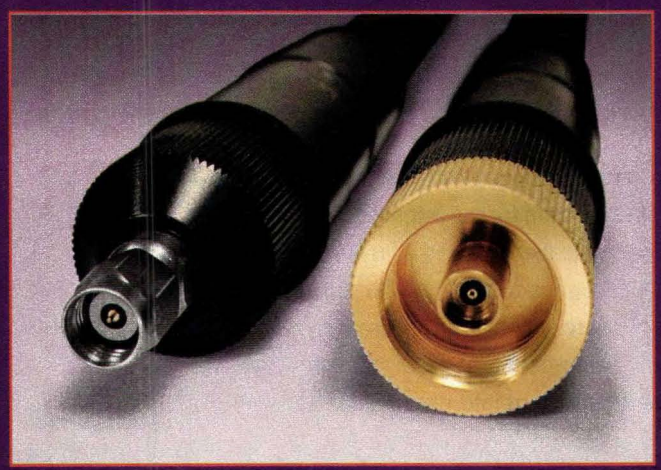

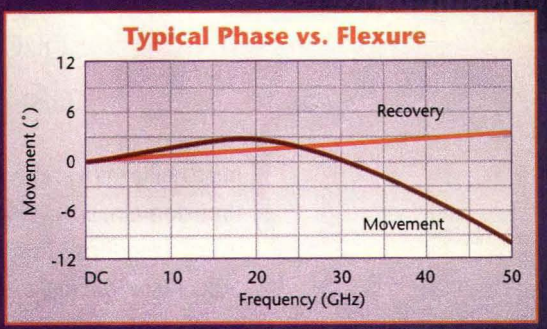

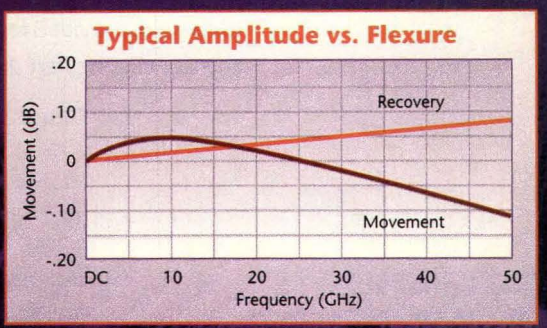

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DAVID CHANDLER
VCXO / Hybrid Design Engineer
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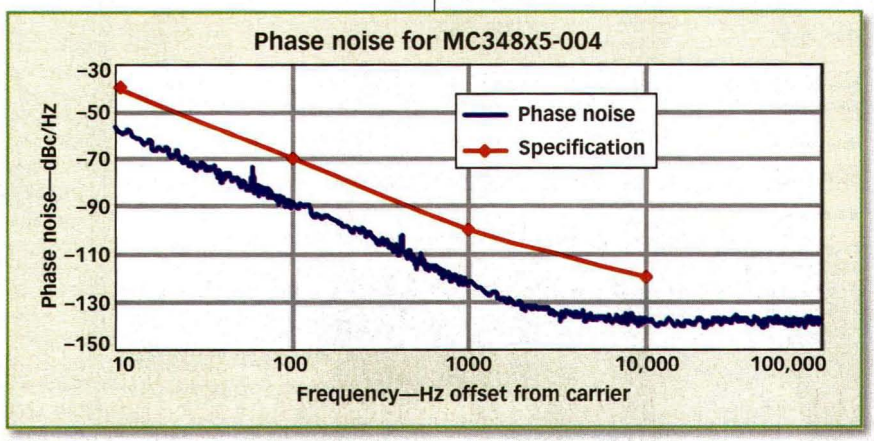
Figure 9 is a phase-noise plot for a 155.52-MHz, +3.3-VDC LVPECL VCXO. Noise floors of commercial phase-noise equipment are as low as −170 dBc/Hz. For several communications applications, only the noise at certain frequencies away from the carrier is a concern, so the phase noise is specified at these frequencies. For many applications, the total noise power over a specific range of frequencies, and not the shape (Fig. 9), is the primary concern. To determine the power over the frequency range (bandwidth), the time-domain signal must be analyzed in the frequency domain, and then reconstructed in the time domain into an RMS value with the unwanted frequencies excluded. This may be done by converting L(f) back to $S_\phi(f)$ over the bandwidth in question, integrating and performing some additional calculations. The result is an RMS value for $\phi(t)$. The result may be expressed in decibels, radians, UI, or seconds. The value, in



9. This phase-noise plot is for a 155.52-MHz +3.3-VDC LVPECL VCXO. The measured performance exceeds the manufacturer's specification.

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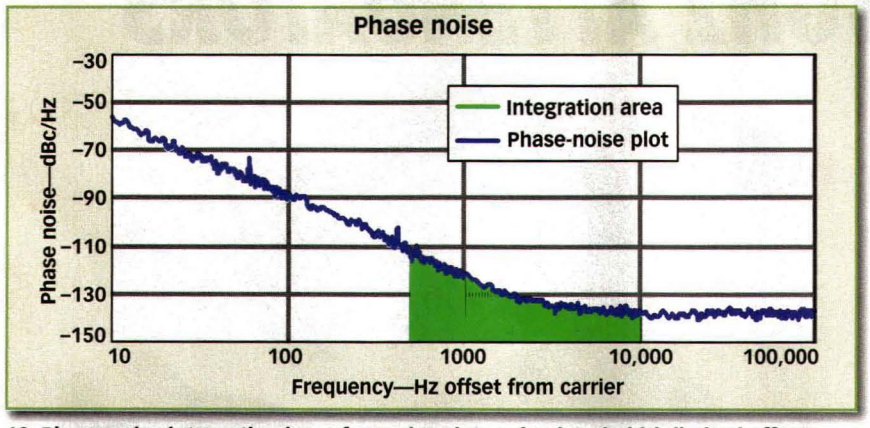

10. Phase-noise integration is performed to determine bandwidth-limited effects (here from 500 Hz to 10 kHz) on phase jitter.

seconds, represents one standard deviation of phase jitter contributed by the noise in the defined bandwidth. **Figure 10** displays the integrated area (in green).

Several telecommunications standards differentiate between jitter above and below 10 Hz, identifying jitter below 10 Hz as "wander." Trace integration can be performed using phase-noise test equipment, or by exporting data from a phase- noise tester into a spreadsheet program. This may seem cumbersome, but it is one of the most-effective ways to accurately provide an RMS value for $\phi(t)$ over a specified bandwidth.

Only ϕ_{RMS} can be determined in this manner. The peak-to-peak value must be approximated. (These approximations are discussed in Appendix B, which is included in the long version of this article available on the *Microwaves & RF* website at www.mwrf.com.)

Note the order-of-magnitude difference between the integrated phase-noise plot and the time-domain analysis. As explained earlier, there are at least two major causes for these differences:

1. Definition of bandwidth. The bandwidth is well-defined and limited when performing a phase-noise integration. There is no limitation of bandwidth on the oscilloscope method other than the equipment limitations.
2. Limitations imposed by the digital oscilloscope's noise floor (about 2 ps).

The measured values can vary greatly depending on test techniques. For this reason, it is critical for a designer to understand the jitter tolerance for a VCXO in a higher-level system. MRF

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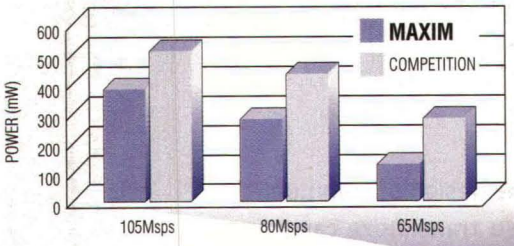

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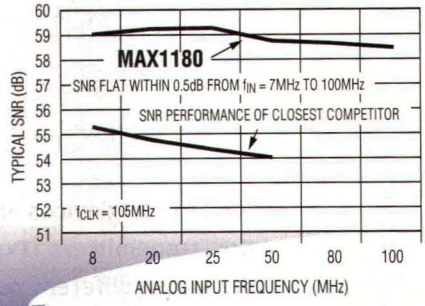

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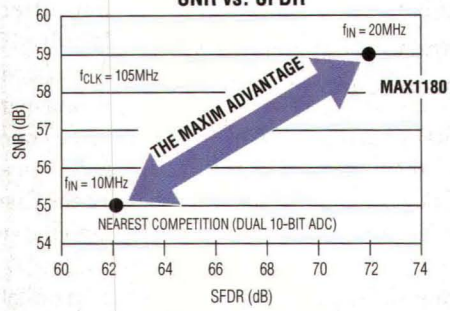

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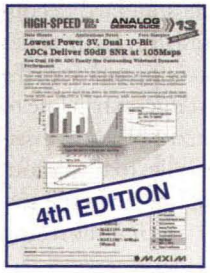

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Setting Bias In High-Power GaAs FET Devices

High-power GaAs transistors can be unstable under certain conditions. As a result, special attention should be paid to the design of biasing and DC blocking circuitry to ensure reliable operation.

Power GaAs FET devices are vital to the design of communications transmitters (Txs), but these transistors can be finicky under different power-supply conditions. Part 1 of this two-part article (see *Microwaves & RF*, January 2002, p. 66) reviewed turn-on and turn-off sequencing, gate-biasing circuitry, and a generic PA for the purpose of explaining different biasing approaches. The conclusion of this article will examine the drain-bias circuitry and DC-blocking elements for a typical power GaAs FET device.

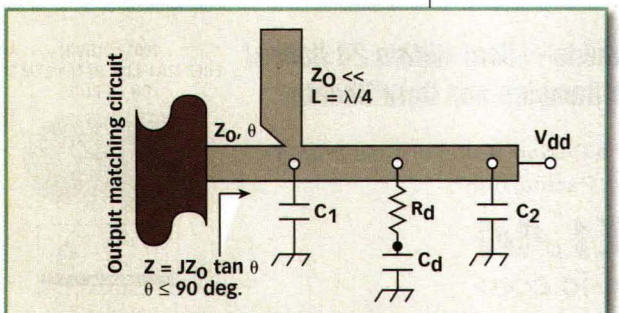
RAYMOND BASSET
Director, Power Product Engineering
Fujitsu Compound Semiconductor, Inc., 2355 Zanker Rd., San Jose, CA 95131; (408) 232-9600, Internet: www.fcsi.fujitsu.com.

Drain-bias circuitry has several functions:

- To maintain a constant drain-to-source voltage.
- To supply drain current of at least I_{dsmax} under drive conditions.
- To stabilize the device outside the amplifier bandwidth.
- To filter the desired RF signal, the spurious products, and harmonics generated by the device from low frequencies to high frequencies. For the input-biasing circuit, the output-biasing circuit can also be used as a matching element.
- To isolate the gate from any signal from the output circuit.

Figure 3 shows a block diagram for an amplifier using the FLL1500IU-2C device, with gate-bias circuit, drain-bias circuit, and matching circuits, while **Fig. 6** shows the drain-bias circuitry. If this latter circuit is used for matching, the length of the first line is shorter than 90 deg. and the circuit presents an impedance of $Z = jZ_o\tan\theta$ (assuming low losses) in parallel with the output-matching circuit. If the function of this circuit is solely biasing, then $\theta = 90$ deg. and its impedance is infinite at the amplifier's fundamental frequency.

In addition to the standard decoupling capacitors, a resistor, R_d, in series with capacitor C_d, is connected from the extremity of the first microstrip line to the ground. This circuit brings a dissipating element in parallel to the decoupling capacitors and improves the stability of the amplifier. The first microstrip line has to carry large drain currents (I_{ds}) of up to 20 A and more for large transistors. This means that its minimum width is limited and consequently its maximum characteristic impedance cannot be high. To minimize the DC drop in the biasing circuit, the microstrip line that is connected to the matching circuit should have a width that is as large



C_1: Capacitor with a series resonant frequency at the amplifier fundamental frequency. That includes the inductance due to the ground via holes
C_2: 100 pF, 1, 10, 100 nF, 1, and 10 μF with low parasitic inductances and resistances
R_d: about 50 Ω
C_d: about 1.0 nF

6. This generic amplifier drain-biasing circuit can also provide output matching.

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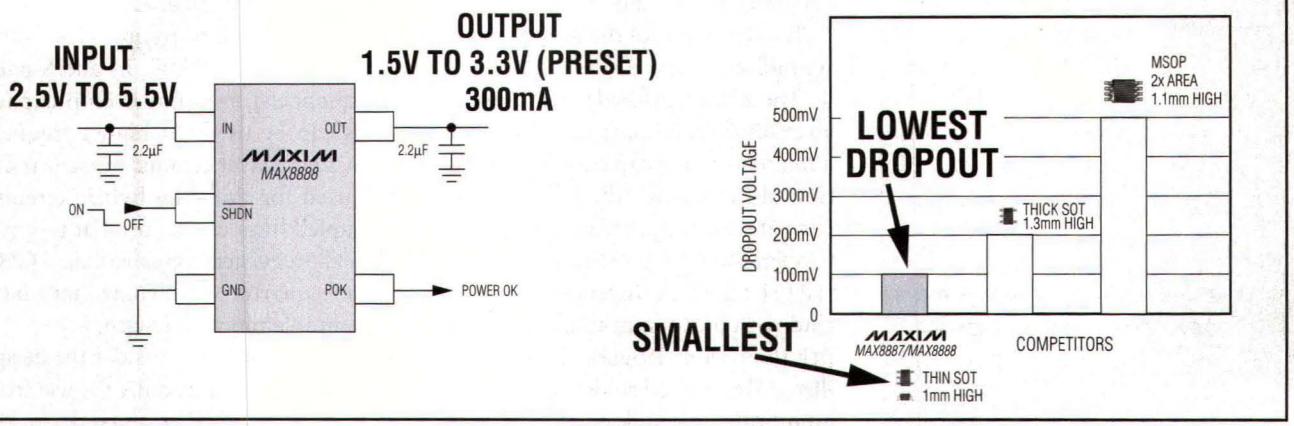

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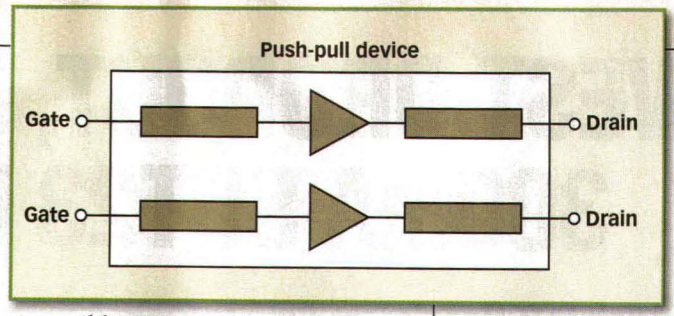

7. This push-pull device uses no internal transversal links.

as possible. However, its characteristic impedance must be compatible with the amplifier bandwidth and this line should provide a cutoff frequency for the amplifier's first higher-order mode, which is approximately:

$F_c = 300/[Er^{0.5}(2W + 0.8h)]$

where:

F_c = the cutoff frequency (in gigahertz),

W = the width of the microstrip line (in millimeters), and

h = the height of the microstrip line (in millimeters).

The table mentioned in ref. 1 provides an example of the maximum current that a microstrip line can handle as a function of the line width. This is the DC current that is required to raise the trace temperature 100°C on microstrip material [31-mil-thick Rogers 4350 material clad with 0.068-mm-thick, 2-oz. copper (Cu) from Rogers Corp. (Chandler, AZ)]. A good solder connection to an infinite heat sink on the bottom of the circuit-board material is assumed.

For a model FLL1500IU-2C device with a total maximum current of 30 A and assuming that each side of the device has a bias circuit, the minimum line width should be 1 mm (15 A for each side). To minimize the DC-voltage drop and have some operating margin, a 1.5-mm line should be used. The matching circuit between the bias circuit and the device drain connection should use lines with a wider width since it must carry DC and RF currents.

Capacitor C1 in Fig. 6 is not needed when an open-circuited quarter-wavelength low-impedance line is used since its input impedance is null at the fundamental frequency of operation.

Instead of using a low-impedance quarter-wavelength open-circuited shunt stub in the gate- and drain-bias circuits to realize an RF short circuit, a radial shunt stub[2,3] can be used. It creates a low impedance to the ground at a precise point and is physically shorter than the equivalent transmission line.

Several types of DC-blocking elements can be used in the input and output circuits. They can also have two functions—DC blocking and RF matching. For the output, they should be selected for low loss and high-power capability. Only the output circuit will be considered here, limited to the case of DC-blocking elements with low reflection inserted in series with the output 50- or 25-Ω transmission line.

At relative low frequencies [ultra-high frequency (UHF), L- and S-band frequencies], multilayer chip capacitors can be used. At higher frequencies, single-layer ceramic capacitors can be used for thin-film hybrid circuits. Coupled lines can be used at relatively high frequencies (greater than 3 GHz) where quarter-wavelength lines have acceptable physical length.

As an example, consider the design of a 150-W balanced PA for use from 2.11 to 2.14 GHz, based on the FLL1500IU-2C (push-pull) device. For L- and S-band applications where plastic circuit boards are used, multilayer chip capacitors in small cases are often mounted in series with the input and output circuits to block the DC.

This capacitor should have an impedance that is low compared to 50 Ω, such as a capacitor with a resonant series frequency near 2.14 GHz.[4] Capacitor ATC 100 A from American Technical Ceramics (Huntington Station, NY), for example, has a value of 20 pF and resonates at 2 GHz. The firm states the capacitor's series resistance as being 0.15 Ω at 1 GHz. The series resistance at 2.14 GHz can be calculated from the following skin-effect formula:

$RSF2 = RSF1(F2/F1)^{0.5} = 0.15(2.14/1.0)^{0.5} = 0.22\ \Omega$

The capacitor's impedance at 2.14 GHz is Z = 0.22 Ω, which is low compared to 50 Ω. Addition of this capacitor will not affect the output matching significantly. Its insertion loss in a 50-Ω line with VSWR of 1.10:1 is:

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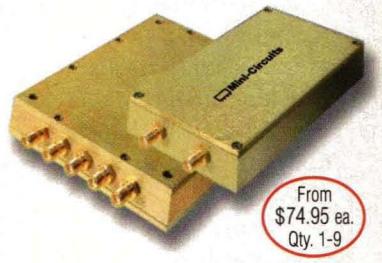

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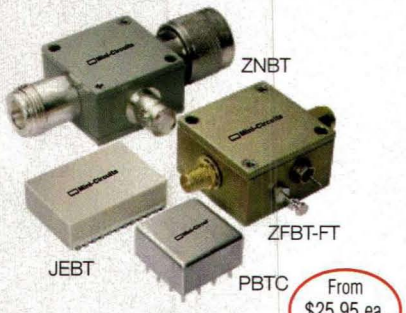

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Loss = 20log[(2R + Rs)/2R]
where:

R = 50 Ω and R_s = 0.22 Ω.

Substituting these values yields a loss of 0.019 dB. The power dissipated in the device can be found from:

$P_{diss} = P_{out}(R_s/R)$
where:

P_{out} = one-half the output power of the push-pull device since each side of the device has a separated bias and matching circuit. In this example, the device is used in a balanced configuration with 50-Ω quadrature couplers:

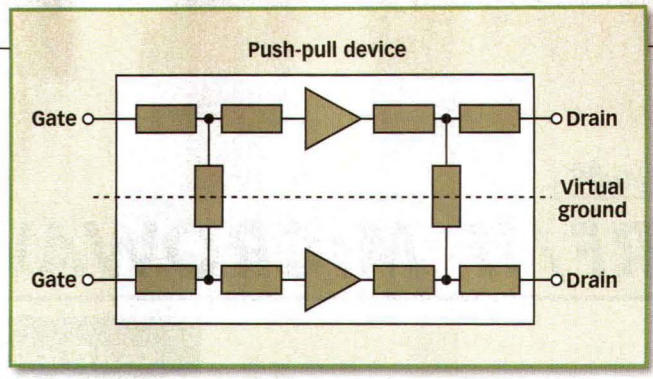
$P_{diss} = (75 \times 0.22)/50 = 0.33$ W

The capacitor manufacturer provides the thermal resistance and the maximum capacitor recommended temperature as being R_{th} = 11.4°C/W and T_{max} = +125°C.

For the example amplifier and a temperature of +70°C, the temperature of the capacitor is T_c = 70 + 11.4(0.33) = +74°C, which is lower than the manufacturer's maximum temperature.

Coupled microstrip transmission lines used as DC blocks are easy to integrate into an amplifier layout, and are a good alternative to lumped capacitors. However, at low frequencies, these quarter-wave lines require too much circuit-board space.[5]

Push-pull devices often have no transversal internal connections **(Fig. 7)**. Both sides of the device are independent and can be combined in as many ways as two single-ended devices. However, real push-pull devices **(Fig. 8)** can be used only in push-pull amplifiers since they have internal transversal connections that create a virtual ground. This article only considers bias circuits that do not use the virtual ground of the push-pull configuration. Note also that each side of a push-pull device has the same bias circuits as a single-ended device **(Fig. 9)**.



8. This push-pull device has internal transversal connections.

This means that a push-pull device has two gate-bias circuits and two drain-bias circuits. Since the chips in a push-pull device are matched, the two gate-bias circuits may be connected in parallel to a single voltage source without creating a quiescent drain-current unbalance between the two sides of the device. The total quiescent drain current is tuned with the single gate-voltage source. The resistance experienced by each

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gate, including the internal resistance of the voltage source R_{IN}, should equal the recommended data-sheet gate resistance: $R_G = R_C + 2R_{IN}$, where R_C is the resistance connected to each gate.

The device's data sheet provides the maximum ratings for a flange temperature of +25°C. The channel temperature (T_{ch}) can be calculated from the device data-sheet thermal resistance using:

$$T_{ch} = T_f + P_{diss}R_{th}$$

where:

T_{ch} = channel temperature (in °C),

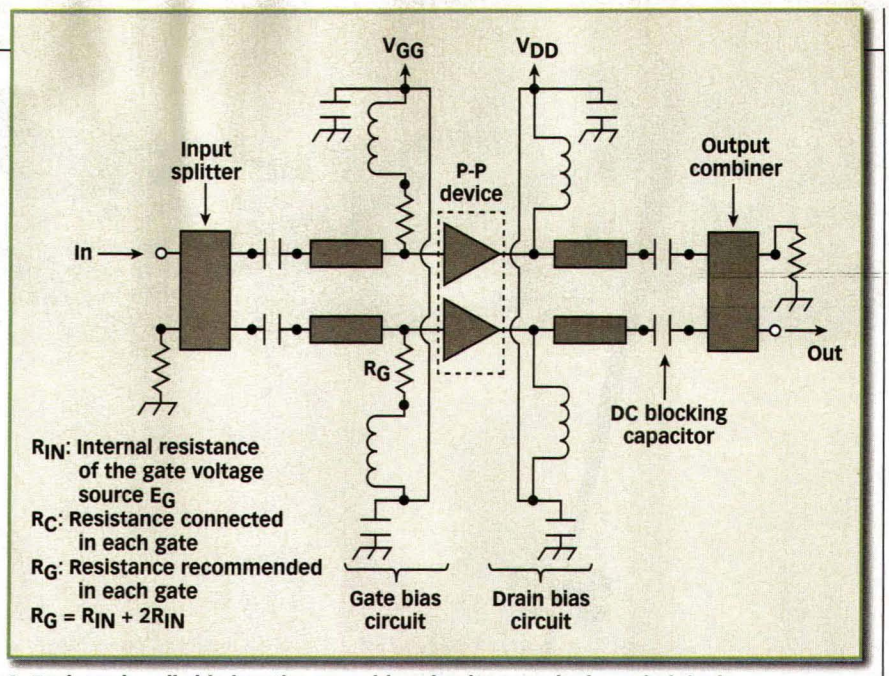
T_f = device flange temperature (in °C),

R_{th} = device thermal resistance (in °C/W), and

P_{diss} = power dissipated in the device (in W).

With RF applied, $P_{diss} = V_{ds}I_{ds} + P_{in} - P_{out}$, but without RF applied, $P_{diss} = V_{ds}I_{dsq}$.

Thermal resistance can also be calculated with the methods of ref. 6. MRF



9. Each push-pull side has the same bias circuits as a single-ended device.

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Noise Figure (dB)*	2.4	2.1 (Cell)/2.2 (PCS)	1.9 (Cell)/2.2 (PCS)	2.0/1.8 (Cell)/2.1 (PCS)/1.8 (GPS)
IIP3 (dBm)	+9 (LNA)	+11 (Cell LNA) +8 (PCS LNA)	-9.0 (Cell)/-8.5 (PCS) Cascaded results for RX path	+11 (Cell LNA)/+8 (PCS LNA)/ +3 (GPS LNA)
Package	4x4mm LCC	5x5mm LCC	8x8mm module	5x5mm LCC
Technology	SiGe HBT	SiGe HBT	SiGe HBT	SiGe HBT
Application	Cell CDMA AMPS	Cell/PCS CDMA AMPS	Cell/PCS CDMA AMPS	Cell/PCS CDMA AMPS GPS

*Cascaded results **Includes integrated RF SAW filters ***Differ in gain control method for IS-98 IMD tests

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Integrated VGA Aids Precise Gain Control

An integrated VGA circuit with fixed-gain amplification and variable attenuation is useful for power leveling in both receivers and transmitters, even at high intermediate frequencies.

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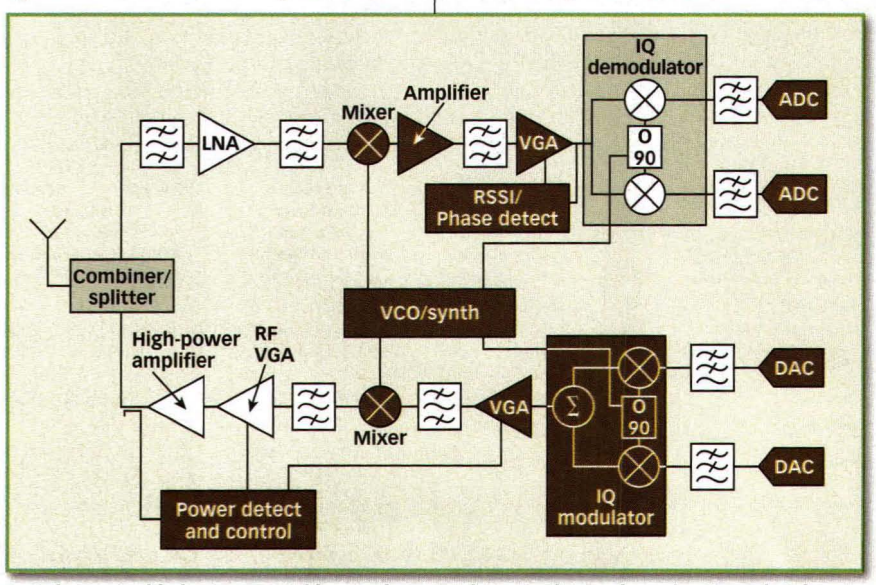
EAMON NASH
Applications Engineer
Analog Devices, 804 Woburn St. MS-122, Wilmington, MA 01887; (781) 937-1239, e-mail: eamon.nash@analog.com.

Gain control is necessary in mobile-communications systems because of the wide range of transmitted and received signals. A base station for Global System for Mobile Communications (GSM), for example, must receive and demodulate signals from about –15 dBm to less than –104 dBm, even when in-band blocking signals can be as large as –13 dBm. The goal of the automatic-gain-control (AGC) circuitry in these systems is to provide a relatively constant input level to the analog-to-digital converter (ADC). Although gain control can be implemented at RF, intermediate frequency (IF), or baseband frequencies, most gain control operates at IF. While a precise relationship between gain control voltage and gain is always desirable, in receiver (Rx) applications the ability to deliver a constant signal level to the subsequent stages is generally more critical. Because of this, an analog or digital level detector is an important component for setting precise gain levels in mobile-communications systems.

Although mobile-communications



1. Gain control is instrumental in receivers and transmitters, in order to maintain precise power levels.

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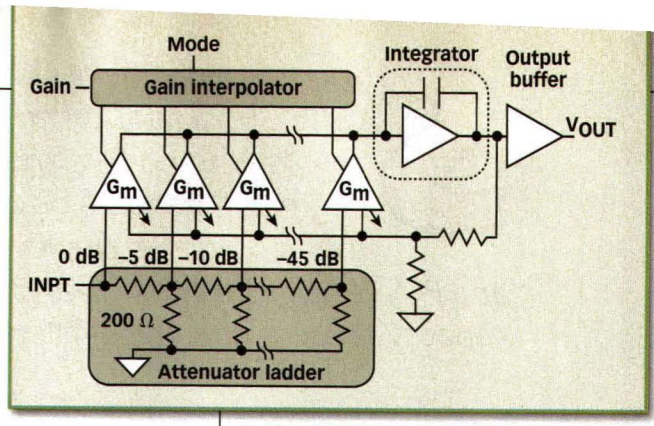
terminals must be able to transmit over a large power range, such as 75 dB in wideband code-division-multiple-access (WCDMA) systems, a large range of power control is also needed in the base station. Network operators must be able to vary cell sizes by fixing the nominal or static base-station power at system installation. In the case of GSM, the nominal base-station power can be adjustable over a



2. The AD8367 X-AMPTM architecture provides precise linear-in-dB gain control over a wide dynamic range.

12-dB range. If the base station has a fixed-gain power amplifer [PA] (which is generally the case), the nominal output of the base station must be set by varying the input signal to the PA **(Fig. 1)**.

In addition, the transmit power will vary depending upon the distance to the mobile unit. Also, time-division-multiple-access (TDMA) systems such as GSM and IS-136 require orderly power ramping. This adds another 40 to 50 dB to the power-control range of the base station. In the case of a CDMA or WCDMA base station, the signal power out must be varied as the call loading in the cell varies.

***Although** mobile-communications terminals must be able to transmit over a large power range, such as 75 dB in WCDMA systems, a large range of power control is also needed in the base station.*

In transmit applications where the output will be constantly varying for any or all of the previously stated reasons, the relationship between the gain of a variable-gain amplifier (VGA) and its gain-control voltage becomes important in the struggle to set the correct power. For example, if the VGA has a gain-control relationship that is linear, temperature-stable, and flat within the band of interest, a simple two-point VGA calibration will be sufficient. Using the calibration data, the base-station controller can confidently set the power

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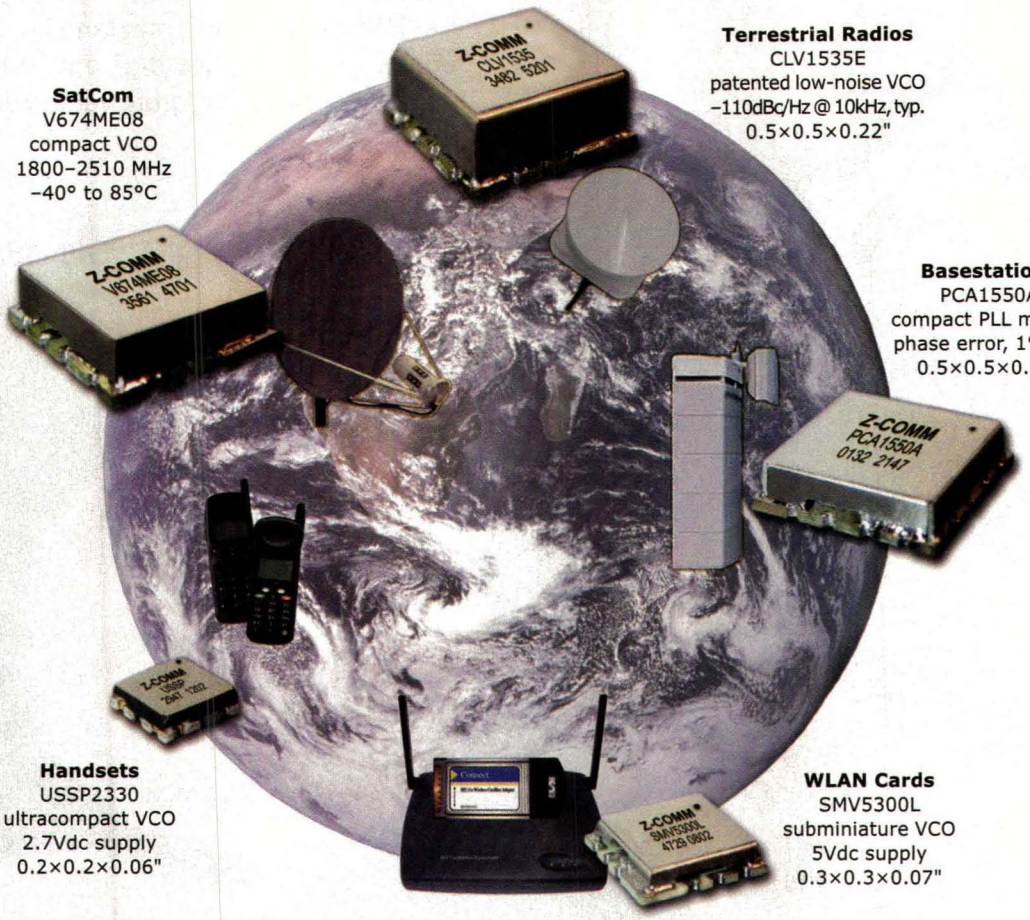

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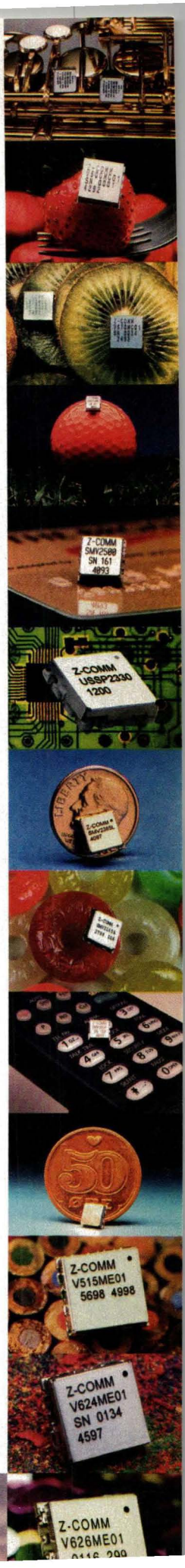

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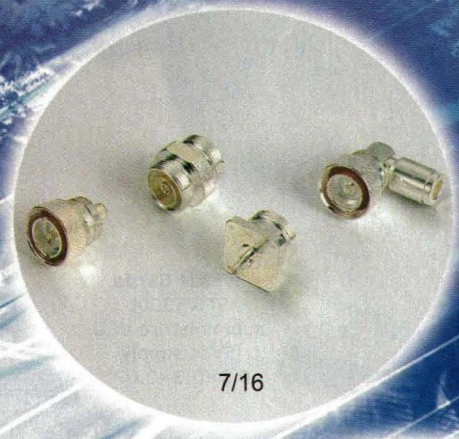

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SMA,SMB,SMC, MCX

1.0/2.3
1.6/5.6

N, BNC, TNC, MINI-UHF

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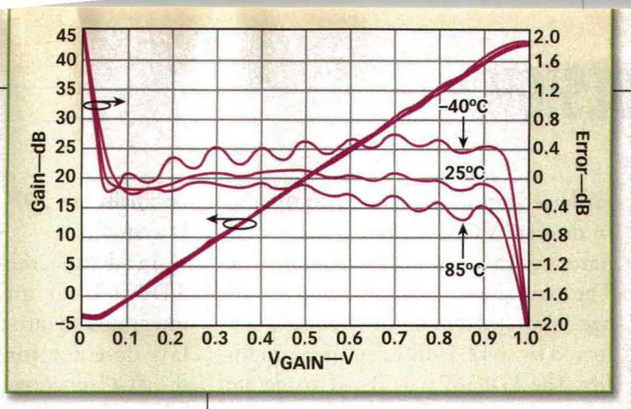
and ignore changes in temperature and frequency. Without such linearity, however, a number of calibration points will be needed across the frequency band of interest. And if the performance varies as a function of temperature, additional calibration points will be needed to account for these temperature-dependent variations of gain-control voltage.

An AGC circuit typically consists of a variable attenuator followed by fixed-gain amplifiers. By integrating these functions on a single chip, it is possible to develop VGAs with high performance levels while dramatically shrinking the overall size of the circuitry. One development goal for a new generation of VGAs from Analog Devices

In transmit applications where the output will be constantly varying, the relationship between the gain of the VGA and its gain-control voltage becomes important in the struggle to set the correct power.

was to meet performance requirements for a variety of cellular base-station systems over a wide range of IFs, even though achieving a temperature-stable, linear-in-dB gain-control range at high frequencies represents a significant challenge. For example, IFs in single-conversion superheterodyne Rxs for cellular base stations can be as high as 380 MHz. One of those integrated VGAs, the model AD8367, integrates a variable atttenuator with a 45-dB range (0-to-45-dB attenuation) with a very linear fixed-gain amplifier **(Fig. 2)**.

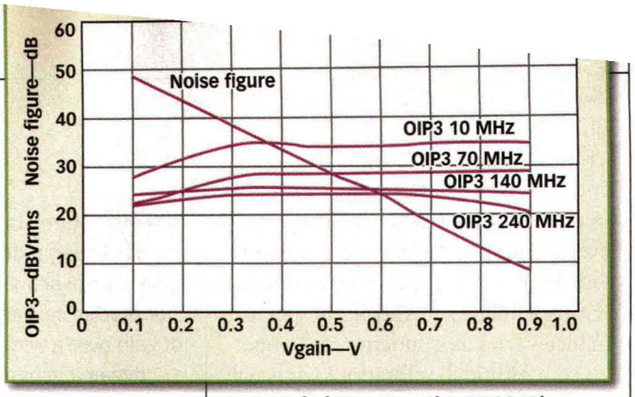
The AD8367 is based on the company's patented X-Amp™ architecture named after the exponential nature of the gain-control function). The precise near-in-dB scaling [i.e., (gain (in dB)/(VAGC (in V) is constant] significantly simplifies AGC design in cellu-



3. Temperature-stable, linear-in-dB gain control allows for precise setting of gain over a wide dynamic range.

lar base stations compared to the use of discrete VGAs or PIN-diode attenuators with cascaded gain amplifiers. The main signal path consists of a voltage-controlled 0-to-45-dB attenuator followed by a 42.5-dB fixed-gain amplifier. The AD8367 is designed to operate optimally in a 200-Ω system.

In addition, the AD8367 has an integrated square-law detector for the AGC function.



4. As gain increases, the AD8367's noise figure improves while the OIP3 remains fairly constant.

In AGC mode, the gain-bias pin provides the received-signal-strength-indication (RSSI) control and the output signal is leveled 354 mV RMS (1 V p-p for an unmodulated sine wave). This circuit is particularly useful if an Rx must handle signals with different modulation formats. A good example of this would be a modern GSM base station that must receive both Gaussian minimum-shift-keying (GMSK) and eight-state phase-shift-keying (8-PSK) modulation (EDGE) signals.

The integration of attenuator followed by fixed-gain amplifier is an ideally suited to Rx AGC circuits, since the input third-order-intercept (IIP3) and the noise figure track dB per dB with the gain setting **(Fig. 3)**. When the input signal is strongest, the gain is at a minimum and the IIP3 and noise figure are at maximum levels. When a received signal is weakest, the gain is at a maximum and the IIP3 and noise figure are optimum for the weak signal, thereby maintaining a wide signal dynamic range. The output third-order-intercept (OIP3) and output power at 1-dB compression are independent of the gain and temperature, which is another significant advantage over discrete VGAs, which can suffer drifting output power levels with temperature and gain settings **(Fig. 4)**.

The AD8367 is packaged in a 14-lead thin shrunk small outline package (TSSOP) and is characterized for operating temperatures from −40 to +85°C It operates on a single voltage supply from +3 to +5 VDC. The integrated device has a 3-dB bandwidth of 50 MHz, and has been thoroughly evaluated at common IFs such as 70, 140, 190, and 240 MHz. Evaluation boards and samples are now available. MRF

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Test designs for Bluetooth qualification

Bluetooth is a wireless technology that enables limited-range RF interconnectivity between information appliances (see the Bluetooth supplement to *Microwaves & RF*, December 2000). A 10-page application note, "Be Ready for *Bluetooth*™" by Isabelle Duverne of Agilent Technologies (Santa Clara, CA), addresses the complexities inherent in testing Bluetooth designs. Consisting of a radio unit, a baseband link-control unit, and link management software, a Bluetooth system is also comprised of higher-level software utilities that concentrate on interoperability and functionality. The lowest layer of the protocol stack is the Bluetooth radio layer, and the Bluetooth Special Interest Group (SIG) has created a list of tests to be performed for its qualification.

Tx testing and RF power-measurement testing should be performed to ensure that transmit power levels can maintain links, minimize interference, and maximize battery life. Transmit output-spectrum measurements test the power levels in the frequency domain to ensure the minimization of out-of-channel emissions, which, in turn, reduces system interference and ensures regulatory compliance. Modulation requirements include initial-carrier frequency tolerance, modulation characteristics, and carrier frequency drift. Timing tests include analyzing the burst profile, PLL, settling time, and other timing characteristics. Out-of-band spurious emissions testing provide verification of the Bluetooth radio's operation within regulatory parameters.

Rx testing ensures that BERs do not rise beyond suitable levels under a number of conditions. Bluetooth Rx sensitivity is a measurement of the minimum signal level that the Rx needs to produce the maximum permitted BER. Sensitivity testing is required for single and multislot packets. Rx measurements covered in the note also include carrier-to-interference performance, blocking performance, and the maximum input-level test. This note is available as a free download from the company's website.

Agilent Technologies, 5301 Stevens Creek Blvd., Santa Clara, CA 95052; (800) 452-4844, (650) 752-5000, FAX: (650) 752-5633, Internet: www.agilent.com/find/bluetooth.

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Tx testing and RF power-measurement testing should be performed to ensure that transmit power levels can maintain links, minimize interference, and maximize battery life.

Understand signal-generator output specifications

Signal generators are complex systems that deliver a user-specified amount of RF power at a selected frequency. They are called upon to operate with or without modulation and with as few artifacts as possible. While the generators are specified to provide the user with a way to verify, real-world applications may require a more thorough understanding of an instrument's performance. An eight-page application note, "Understanding Your Signal Generator Output Specification" by David Owen of IFR (Wichita, KS), examines the building blocks that are used in the output stages of a typical generator and explains how they contribute and interact to level accuracy and IM performance.

When discussing generator output systems, the author points out that basic signal generators are limited to CW and other AM schemes and lack the refinements of ALC and RPP. He lays out a more practical scheme that includes ALC output levelling where an RF drive signal is applied to the output amplifier through a limited-range variable attenuator. In his examination of output impedance, Owen provides a testing method to ensure that the generator's output impedance is 50 Ω. In his analysis of RF-level accuracy, Owen claims that RF-level accuracy cannot be separated from the generator's output VSWR specification.

A comparison of mechanical- and electronic-attenuator implementation for the purpose of determining RF-level accuracy is provided. An explanation of the RPP system and its importance to the design of the output system is included. Methods used to generate AM and to suppress transient behavior are explained. The note offers a test scenario for generators connected together through a combiner. "Understanding Your Signal Generator Output Specification" is available as a free download from the company's website.

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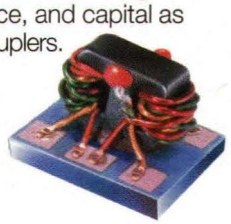

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12dB	DBTC-12-4	5-1000	0.7	21
13dB	DBTC-13-4	5-1000	0.7	18
13dB	DBTC-13-5-75	5-1000	1.0	19
		1000-1500	1.4	17
16dB	DBTC-16-5-75	5-1000	1.0	21
		1000-1500	1.3	19
17dB	DBTC-17-5	50-1000	0.9	20
		1000-1500	1.0	20
		1500-2000	1.1	14
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cover story

DDS Enhances Ovenized Oscillator

This clever blend of analog and digital circuitry results in an OCXO with performance rivaling that of more expensive Rb atomic clocks.

Crystal oscillators have long been known as stable sources of RF energy, though they are incomparable to the performance possible from a rubidium (Rb) frequency standard. For a company such as FEI Communications (Mitchel Field, NY), which manufactures both sources, the challenge was to design a crystal oscillator that could approach the stability of an Rb standard, but without the cost and complexity of the atomic-frequency source. By developing a proprietary double-oven structure, and using a direct-digital synthesizer (DDS) as a form of "correction circuit," the company was able to reach beyond the bounds of conventional oven-controlled-crystal-oscillator (OCXO) design to create a line of crystal oscillators (the models FE-205A, FE-405A, and FE-505A) with Rb-like performance. The sources are initially available in a variety of package styles at frequencies from 5 to 25 MHz, with standard frequencies of 10 and 15 MHz.

The crystal oscillator is a mature technology where a variety of techniques have been refined over the years to improve the basic stability of the active circuitry surrounding a crystal resonator. Temperature-compensation circuits have been developed to adjust for the frequency-shifting effects of changing temperatures, while compact ovens have been constructed to maintain the resonant circuitry within a tightly controlled temperature range to minimize temperature-dependent frequency shifts.

The FE-205A **(Fig. 1)**, FE-405A, and FE-505A **(Fig. 2)** oscillators refine traditional oven-controlled oscillator techniques by using a proprietary double-oven structure to control the temperature seen by the crystal resonator—a special designed, rugged stress-compensated-cut (SC-cut) crystal operated in fifth-overtone mode—and its associated active circuitry **(Fig. 3)**. But these are also the first crystal oscillators to incorporate an additional oscillator—a monolithic, direct-digital numerically controlled oscillator (NCO)—to achieve digital adjustments to the crystal-oscillator output frequency. Using a least-significant bit (LSB) of 1.7×10^{-14} as a tuning increment, these new

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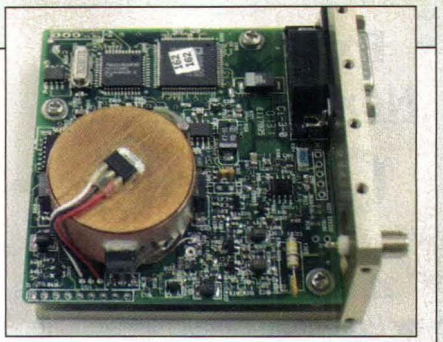
FEI Communications, Inc., A Subsidiary of Frequency Electronics, Inc., 55 Charles Lindbergh Blvd., Mitchel Field, NY 11553; (516) 794-4500, FAX: (516) 794-4340, Internet: www.frequencyelectronics.com.

1. The model FE-205A packs a double-oven SC-cut crystal oscillator with digital frequency correction into a housing measuring only 2.0 × 2.0 × 1.5 in. (5.08 × 5.08 × 3.81 cm).

2. The model FE-405A (left) measures 3.01 × 3.03 × 1.44 in. (7.65 × 7.70 × 3.66 cm), while the model FE-505A (right) measures 2.98 × 2.80 × 0.89 in. (7.57 × 7.11 × 2.26 cm).

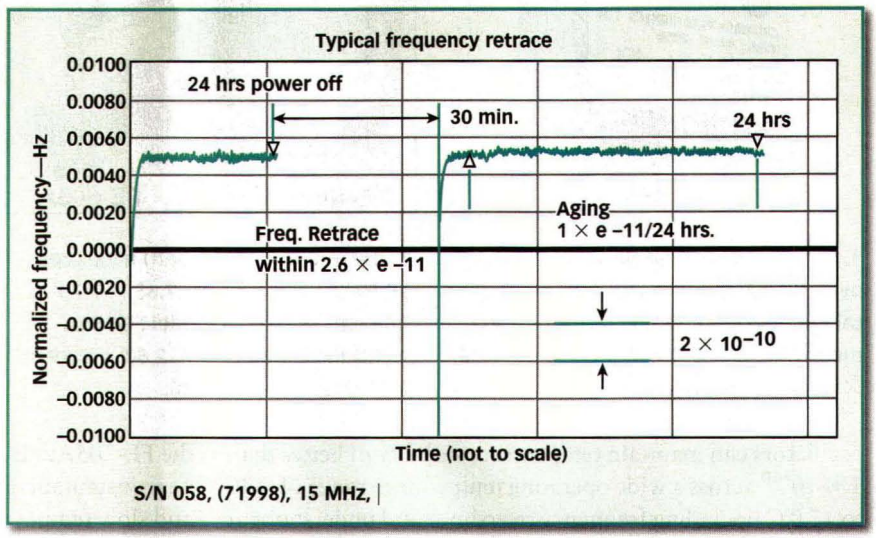


3. This inside look at a model FE-205A crystal oscillator shows the circular double-oven structure to the left and some of the digital ICs used to implement the DDS circuitry.

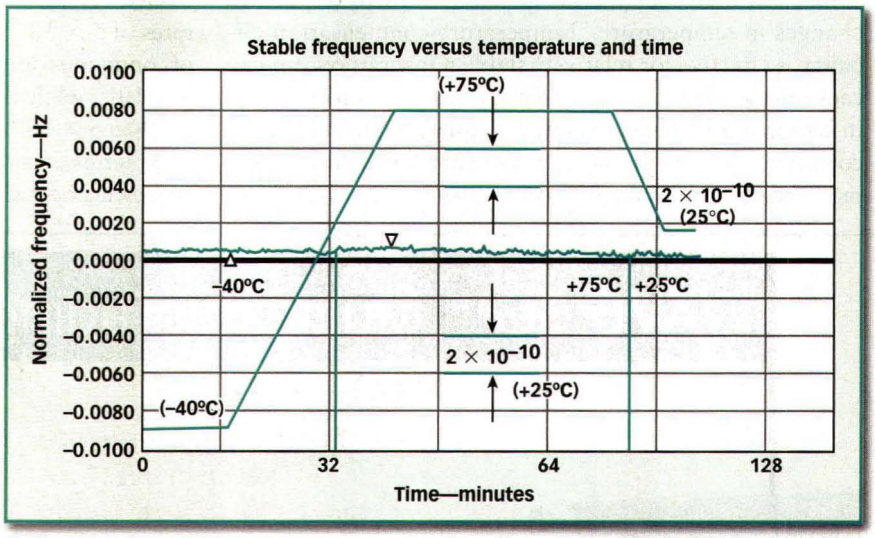
oscillators can maintain temperature stability of better than 1×10^{-10} across a wide operating temperature range of −40 to +75°C (including frequency overshoot and undershoot at any fast or slow rate of temperature change).

Crystal oscillators employing traditional temperature-compensation techniques are generally slow to respond to changes in temperature. Temperature-compensation circuitry is effective for relatively stable temperatures but can cause an overshoot and undershoot with ramps up and down in temperature. The advantage of using digital frequency correction through the incorporation of a DDS is the switching speed of an NCO, on the order of microseconds. This allows the FE-205A, FE-405A, and FE-505A oscillators to achieve near-instantaneous corrections of frequency even with fast and slow temperature slew rates. The new sources adjust so quickly to changes in temperature that they achieve temperature tracking of 5°C/minute. These digital oven-controlled crystal oscillators (DOCXOs) boast typical aging rates of 5×10^{-11} and better than 1×10^{-10} after 14 days of continuous operation. The aging rate per year is better than 1×10^{-8}, while the aging rate over a ten-year period is better than 5×10^{-8}.

In terms of short-term stability, the performance of these new oscillators is similar to that of a Rb standard. The Rb source



4. The typical retrace performance has been measured by shutting off an oscillator for 24 hours and then allowing the source to come back to specified frequency.



5. The true test of the FE-205A Series crystal oscillator is its excellent frequency stability even with rapid ramps up and down in temperature.

provides performance advantages for those applications requiring better long-term stability. Still, the FE-205A, FE-405A, and FE-505A oscillators are designed with approximately one-tenth the total number of parts of a Rb standard, which should translate into a longer mean time before failure (MTBF) for the new oscillators compared to Rb clocks. The short-term stability of the new oscillators is 1×10^{-11} measured over a 1-s period and 2×10^{-12} measured over a 10-s period. The new oscillators require approximately 20 min. warmup time to achieve specified frequency. The retrace performance—the ability to return to frequency after being shut off and turned on again—is 1×10^{-10} in one hour after 24 hours power off, and 5×10^{-10} in 20 min. after 24 hours power off **(Fig. 4)**.

What may be the most challenging test of performance for any crystal oscillator is an evaluation of frequency stability with rapid increases (ramp up) and decreases (ramp down) in temperature. The stability of the FE-205A Series oscillators was measured and plotted normalized to 2×10^{-10} over a total time span of slightly more than 128 min. During that time, the temperature was rapidly raised from −40 to +75°C, and then dropped rapidly to +25°C. As the measurements show **(Fig. 5)**, the frequency wavered very little from the nominal 2×10^{-10}.

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TYPICAL SPECIFICATIONS AT 25°C:

Model	Freq.■ (MHz)	Gain (dB) 0.1GHz	Gain (dB) 2GHz	Flatness† DC-2GHz (dB)	Max. Power Out▲ @1dB Comp. (dBm)	Dynamic Range▲ NF (dB)	Dynamic Range▲ IP3 (dBm)	Thermal Resist. θjc, °C/W	DC Operating Current (mA)	Power Volt	Price $ea. (25 Qty.)
Gali 1	DC-8000	12.7	11.8	±0.5	12.2	4.5	27	108	40	3.4	.99
Gali 21	DC-8000	14.3	13.1	±0.6	12.6	4.0	27	128	40	3.5	.99
Gali 2	DC-8000	16.2	14.8	±0.7	12.9	4.6	27	101	40	3.5	.99
Gali 33	DC-4000	19.3	17.5	±0.9	13.4	3.9	28	110	40	4.3	.99
Gali 3	DC-3000	22.4	19.1	±1.7	12.5	3.5	25	127	35	3.3	.99
• Gali 6F	DC-4000	12.1	11.6	±0.3	15.8	4.5	35.5	93	50	4.8	1.29
• Gali 4F	DC-4000	14.3	13.4	±0.5	15.3	4.0	32	93	50	4.4	1.29
• Gali 51F	DC-4000	18.0	15.9	±1.0	15.9	3.5	32	78	50	4.4	1.29
• Gali 5F	DC-4000	20.4	17.4	±1.5	15.7	3.5	31.5	103	50	4.3	1.29
• Gali 55	DC-4000	21.9	18.5	±1.7	15.0	3.3	28.5	100	50	4.3	1.29
• Gali 52	DC-2000	22.9	17.8	±2.5	15.5	2.7	32	85	50	4.4	1.29
• Gali S66	DC-3000	22	17.3	±2.4	2.8	2.7	18	136	16	3.5	.99
Gali 6	DC-4000	12.2	11.8	±0.3	18.2	4.5	35.5	93	70	5.0	1.49
Gali 4	DC-4000	14.4	13.5	±0.5	17.5	4.0	34	93	65	4.6	1.49
Gali 51	DC-4000	18.1	16.1	±1.0	18.0	3.5	35	78	65	4.5	1.49
Gali 5	DC-4000	20.6	17.5	±1.6	18.0	3.5	35	103	65	4.4	1.49

■ Low freq. cutoff determined by external coupling capacitors. † Measured in test fixture P/N 90-6-20-26.
▲ Models tested at 2GHz except Gali 4, 5, 6, 51, 52, 6F, 4F, 51F, 5F at 1GHz.
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tal oscillators also compares favorably with that of a Rb standard. Maximum harmonics, for example, are −40 dBc for any of the three models, with spurious levels specified at −65 dBc to 3 GHz. The phase noise for the new oscillators is also impressive, at −85 dBc/Hz offset 1 Hz from the carrier, −95 dBc/Hz offset 10 Hz from the carrier, −125 dBc/Hz offset 100 Hz from the carrier, −135 dBc/Hz offset 1 kHz from the carrier, and −145 dBc/Hz offset 10 kHz from the carrier.

These DOCXOs allow large frequency adjustments with 1-percent linearity, depending on the option. With Option 30, analog-frequency adjustments can be made using input voltages from 0 to +10 VDC over a coarse adjustment range of $\pm 2.4 \times 10^{-7}$ and a fine adjustment range of $\pm 0.5 \times 10^{-8}$. With Option 2, adjustments in frequency can be made through a TTL serial-port interface using an LSB of 1.7×10^{-14}. The adjustment range is ±20 Hz for oscillators with a 15-MHz output frequency and ±9.5 Hz for oscillators with a 10-MHz output frequency. Other trim-range options are available.

The new oscillators deliver sinewave output signals at +12 dBm (±2 dB) into a 50-Ω load. They are suitable for frequency reference use in cellular base stations, Stratum clocks, and in Global Positioning System (GPS) applications. They consume a maximum of 15 W during warmup and only 3.5 W power during steady-state operation at +25°C. The standard power supplies are +15 VDC at 1 A maximum and +5 VDC at 200 mA. As an option, the sources can be specified to run from a single +15-VDC supply. The model FE-205A measures 2.0 × 2.0 × 1.5 in. (5.08 × 5.08 × 3.81 cm). The model FE-405A measures 3.01 × 3.03 × 1.44 in. (7.65 × 7.70 × 3.66 cm), and the model FE-505A measures 2.98 × 2.80 × 0.89 in. (7.57 × 7.11 × 2.26 cm). **FEI Communications, Inc., A Subsidiary of Frequency Electronics, Inc., 55 Charles Lindbergh Blvd., Mitchel Field, NY 11553; (516) 794-4500, FAX: (516) 794-4340, Internet: www.frequency-electronics.com.**

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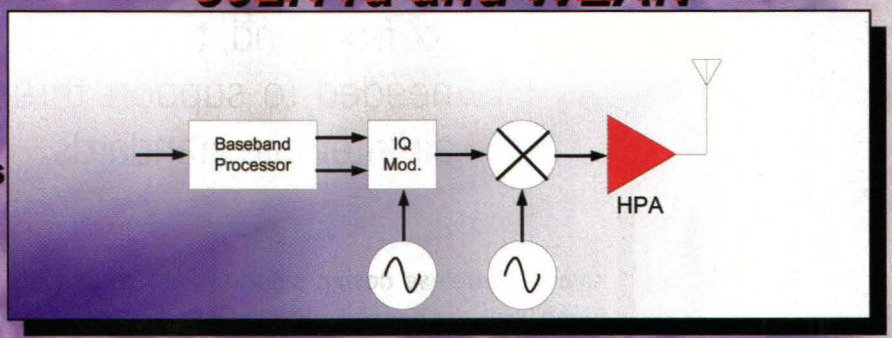

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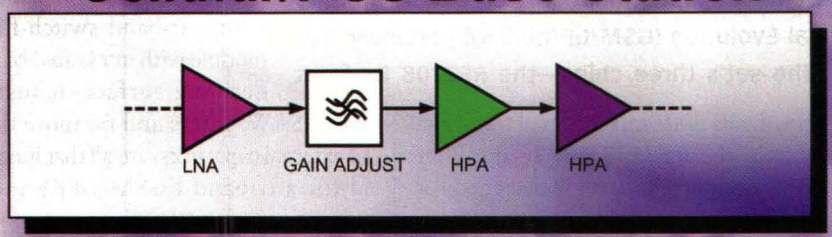

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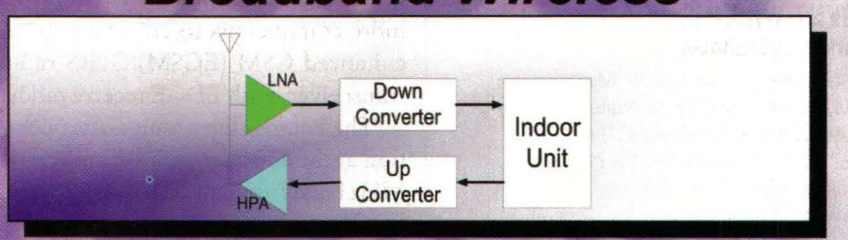

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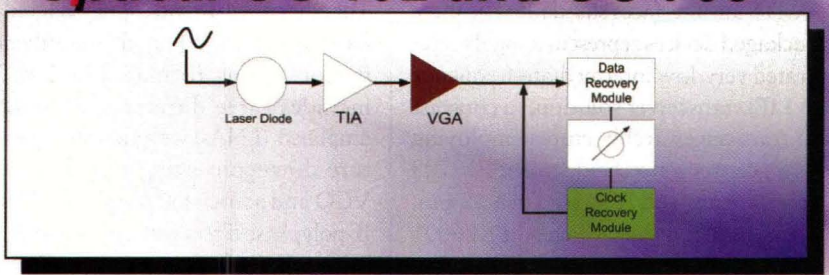

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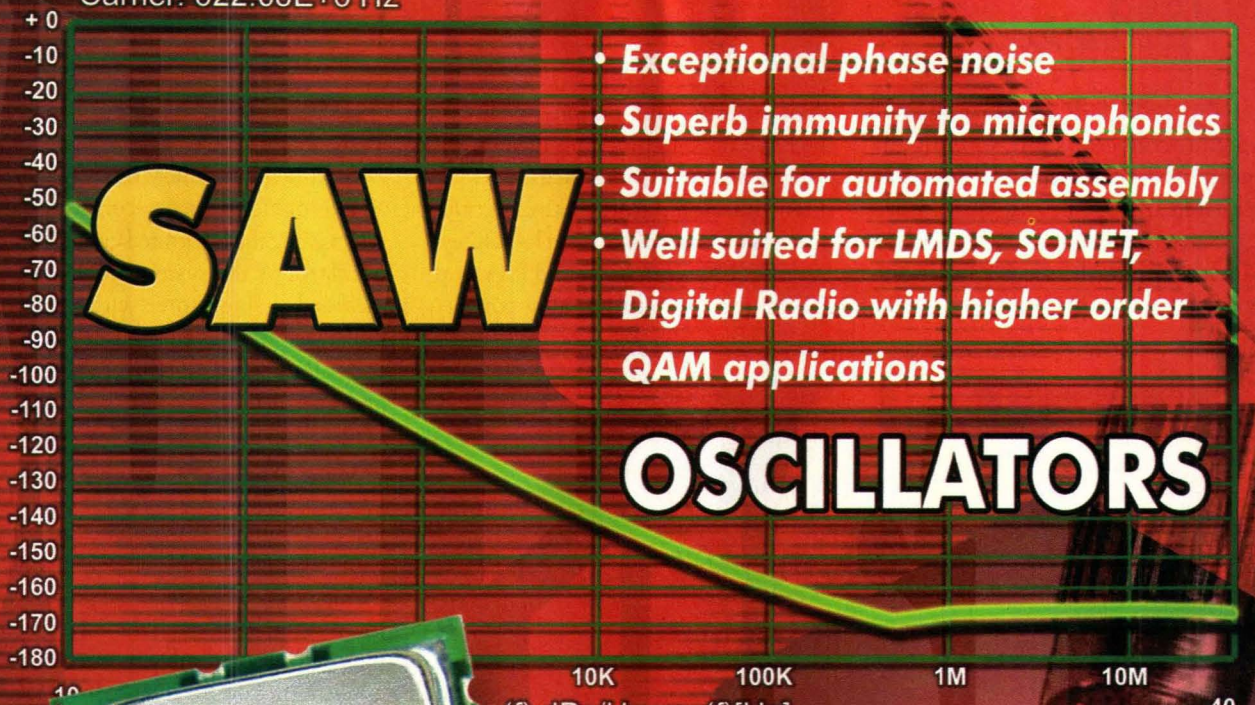
Universal wireless design solutions are sought but rarely found. Still, the new Polaris™ Total Radio™ chip set from RF Micro Devices (Greensboro, NC) may be the nearest thing to a universal solution for Global System for Mobile Communications/General Packet Radio Service/Enhanced Data Rates for Global Evolution (GSM/GPRS/EDGE) receiver (Rx) transceivers. The set's three chips—the RF2708 Rx front end, the RF6001 mixed-signal transceiver, and the RF3133 power-amplifier (PA) module (Fig. 1)—are designed for use with all available baseband controllers supporting GSM, GPRS, and EDGE standards and only require 18 more components to create a triband enhanced GSM (EGSM)/GPRS radio transceiver, with EDGE receive mode.

The Polaris chip set supports operation at the 850/900-, 1800-, and 1900-MHz bands using the GSM, GPRS, and EDGE-receive air-interface standards. The chip set will also be the basis for future chip sets in support of the emerging UMTS air-interfaces standard. The three packaged devices represent a highly integrated very-low intermediate-frequency (VLIF) transceiver solution, in contrast to transceiver architectures employing direct-conversion techniques. The RF2708 receive front-end is housed in a 32-pin, 5 × 5-mm leadless package; the RF6001 mixed-signal transceiver is supplied in a 48-pin, 7 × 7-mm leadless package; and the RF3133 PA module is supplied in a 12-pin, 7 × 10-mm leadless package.

These three devices, along with a triband switch-filter module with integrated band-defining surface-acoustic-wave (SAW) filters and no more than 14 external capacitors, are all that is needed for a triband EGSM/GPRS radio transceiver with EDGE receive mode. The low component count is due to several factors, including use of the VLIF design to eliminate external filters, integration of all voltage-controlled oscillators (VCOs) and associated loop filters, integration of most power-supply decoupling capacitors, and use of a fractional-N Gaussian minimum-shift-keying (GMSK) modulator.

The Rx section is comprised of the RF2708 and portions of the RF6001. Together, they provide a fully filtered signal to the baseband circuitry in either digital or analog format. The RF2708 includes three differential low-noise amplifiers (LNAs) with gain step, quadrature downconverting mixers, receive VCO and associated frequency divider, IF polyphase filters with gain control, and coarse DC-offset-correction circuitry. The RF6001 includes a single frequency synthesizer used for receive and transmit modes. Additional functions for the



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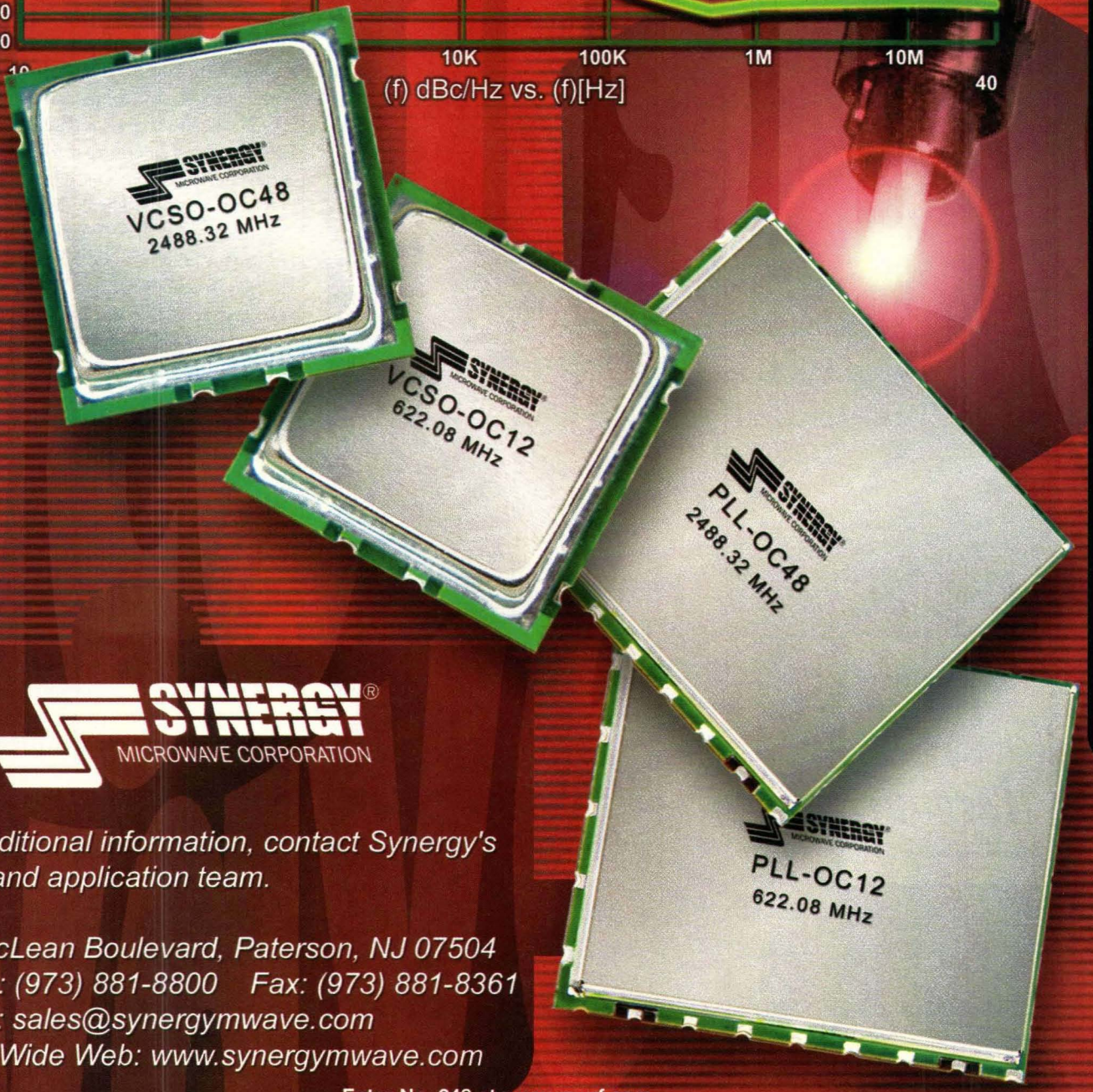

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receive mode include dual 13-b analog-to-digital converters (ADCs), digital downconversion, digital filtering, and digital and analog in-phase/quadrature (I/Q) output interfaces.

In addition to the receive functions mentioned earlier, the RF6001 includes all of the transmit functions, excluding the PA. Transmit functions include digital and analog I/Q interfaces, a GMSK filter and modulator, dual integrated power VCOs and associated loop filters, fractional-N synthesizer and PA ramping-control digital-to-analog converter (DAC). Various regulators and voltage reference circuits, chip or system oscillator, and an auxiliary DAC for use in an automatic-frequency-control scheme when using the integrated circuit's (IC's) oscillator as the system clock are also included. A single three-wire serial data interface provides configuration, mode control, and power down for the RF2708 and RF6001.

The RF3133 PA is a triband, high-efficiency module with integrated indirect closed-loop power-control circuitry. The module is internally matched to 50 Ω, eliminating the need for external matching elements. The integrated power-control function eliminates the need for couplers, detector diodes, power-control application-specific ICs (ASICs), and/or other power-control functions.

The Polaris chip set is based on a VLIF architecture, in contrast to the direct-conversion approach used in some wireless applications. The benefits of VLIF include improved performance and ease of use for GSM and GPRS designs, due largely to the fact that VLIF systems are less sensitive than direct-conversion systems to DC offsets and even-order distortion products. In VLIF designs, any mixer-generated DC components fall outside of the desired passband.

Direct-conversion radios, on the other hand, are more sensitive to DC components since they fall into the passband of interest. The input second-order intercept point (IP2) requirements are higher for direct-conversion radios compared to VLIF designs, since amplitude-modulation (AM) components can also land in the passband of interest. Typically, direct-conversion radios require increased circuit complexity (and power consumption) to combat these problems. Filtering is simpler in VLIF radios, and the input IP2 requirements are reduced (about 20 dB less than for direct conversion) since undesirable AM components are at DC and the desired signal is offset by the VLIF. A polyphase filter is implemented in the RF2708 to keep adjacent and alternate-channel components from being mirrored into the desired signal, meeting the GSM rejection requirements of

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ELECTRICAL SPECIFICATIONS

MODEL	FROM	TO	FREQ RANGE (GHz)	VSWR (GHz)
8006E1	QT3.5mm™ (m) with no nut	3.5mm (f)	DC — 26.5**	DC — 16.0, 1.05 16.0 — 26.5, 1.08
8006E11	QT3.5mm™ (m) with 3/8" dia. nut	3.5mm (f)		
8006E21	QT3.5mm™ (m) with 9/16" dia. nut	3.5mm (f)		
8006Q1	QT3.5mm™ (m) with guide sleeve	3.5mm (f)		

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REPEATABILITY	DC — 18.0 GHz	18.0 — 26.5 GHz
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ELECTRICAL SPECIFICATIONS

MODEL	FROM	TO	FREQ RANGE & MAX. VSWR
7921A	2.4mm Q (f)	2.4mm Q (f)	DC — 26.5 GHz, 1.06 26.5 — 40.0 GHz, 1.10 26.5 — 34.0 GHz, 1.15
7921B	2.4mm Q (f)	2.4mm Q (m)	
7921C	2.4mm Q (f)	2.4mm Q (m)	
8714A1	2.92mm K (f)	2.92mm K (f)	DC — 4.0 GHz, 1.05 4.0 — 20.0 GHz, 1.08 20.0 — 40.0 GHz, 1.12
8714B1	2.92mm K (m)	2.92mm K (m)	
8714C1	2.92mm K (f)	2.92mm K (m)	
8021A2	3.5mm (f)	3.5mm (f)	DC — 18.0 GHz, 1.05 18.0 — 26.5 GHz, 1.08 26.5 — 34.0 GHz, 1.12
8021B2	3.5mm (m)	3.5mm (m)	
8021C2	3.5mm (f)	3.5mm (m)	

Between-Series configurations include: • *2.4mm to 2.92mm (K)* • *2.4mm to 3.5mm*

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European Telecommunications Standards Institute (ETSI) GSM 05.05 v7.1.0 subclause 6.3.

The RF2708 includes three LNAs, each internally matched to 100 Ω differential, supporting a variety of wireless bands including US cellular and PCS, DCS, and EGSM without additional components. A quadband version will be available in the near future.

The fully integrated VCO requires no external support components and works in conjunction with the RF6001, which includes the integrated loop filter and programmable frequency synthesizer. The VCO operates at high-band frequencies; division by two achieves the low-band frequencies.

Quadrature mixers downconvert signals to a 100-kHz IF, which is processed by a three-stage polyphase filter to reject interference and protect the dynamic range of the RF6001's ADCs. Each of the three polyphase-filter stages provides gain control as defined in the gain-control registers set through the serial data interface.

The RF2708 includes a coarse DC-offset-correction scheme to reduce offsets caused by various sources, including device mismatches and local-oscillator (LO) feedthrough. The DC-offset levels are brought to below 10 mV in the RF2708 and are reduced to approximately 50 μV in the RF6001.

Four general-purpose I/Os in the RF2708 enable control of various switch-filter modules transmitter (Tx)/Rx and band-select control pins and aid in coarse DC-offset processing. The logic states of these I/Os are defined through the serial-data-interface registers. By shifting the switchplexer to transmit mode, offsets caused by the RF2708 and associated layout are nullified. Having these I/Os reduces the number of lines between the baseband controller and the radio.

The chip set's receive path includes dual wide-dynamic-range [12.9-b effective number of bits (ENOB)] ADCs at the inputs to the RF6001. The ADCs provide several benefits to the system, including reduced automatic-gain-control (AGC) writes from the baseband device, simplifying operation and the amount of associated software. For baseband devices that can accommodate signals from a few millivolts to approximately +1 VDC, nominal gain is held from reference sensitivity of −109 to −57 dBm before the gain is modified. The generous headroom of these ADCs supports fast up and down fades and the ability to handle large down fades.

The RF2708 provides 70-dB of analog gain control. The LNAs provide a 14-dB gain step. A digital gain-control register provides an additional 90 dB of gain control in 6-dB increments. This much gain control allows a designer to maintain constant digital or analog out-

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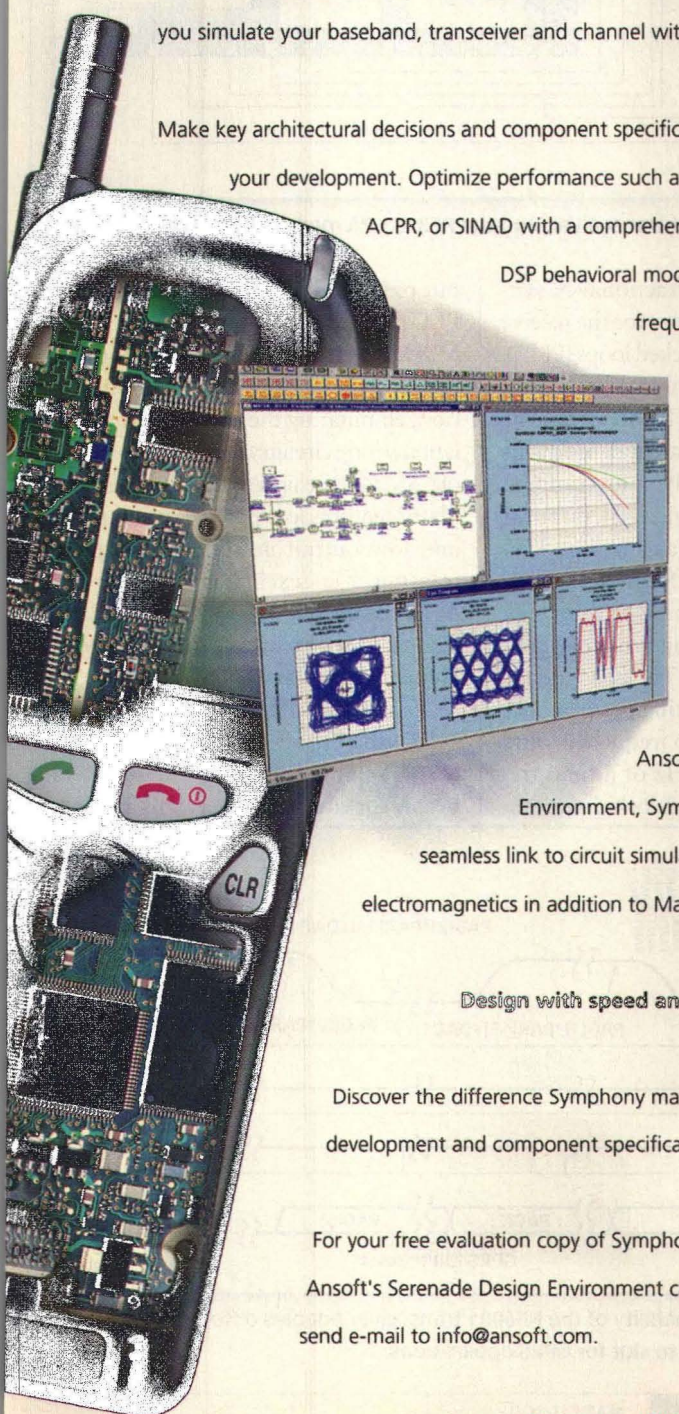

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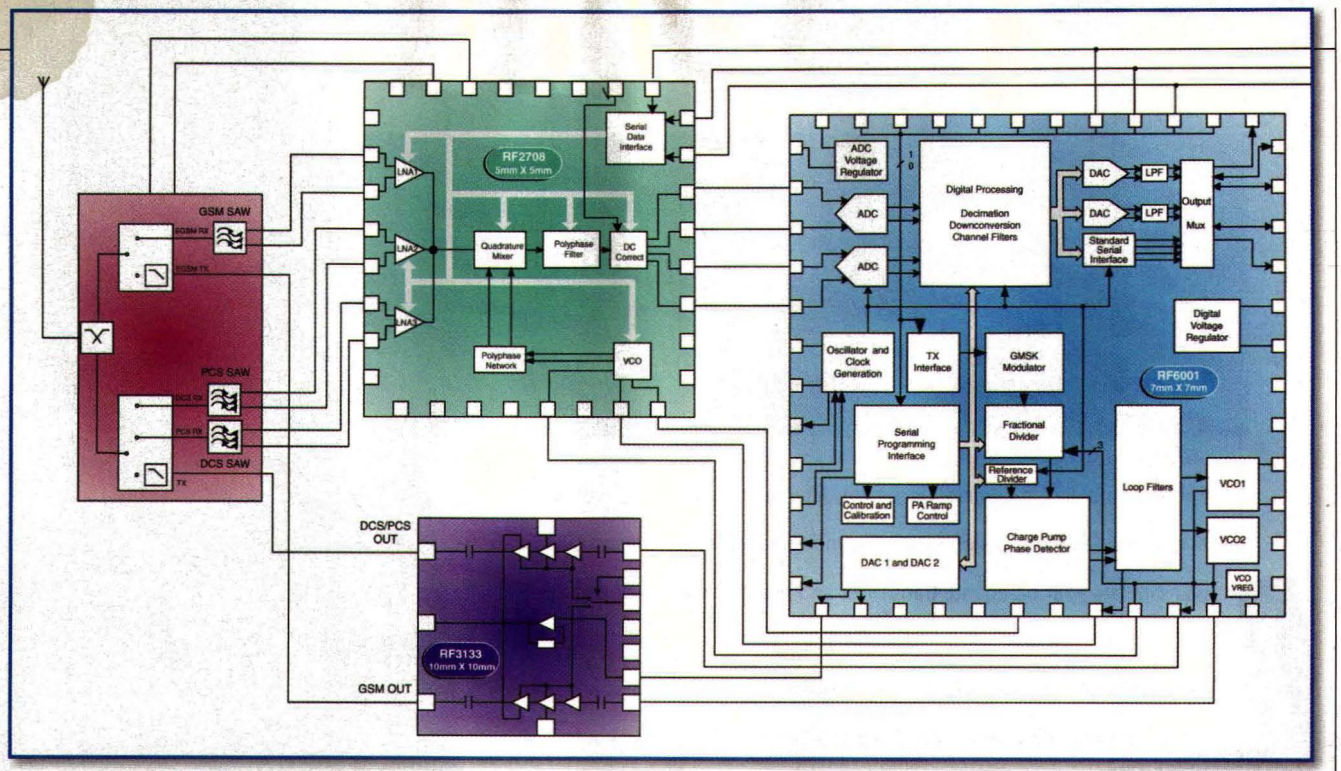


1. The Polaris chip set includes the RF2708 RF front end, the RF6001 transceiver, and the RF3133 PA module.

put voltage to accommodate baseband devices with limited-receive dynamic range. All gain control is performed through the three-wire serial data interface, and all filtering occurs within the Polaris chips to eliminate dependencies on a specific baseband device.

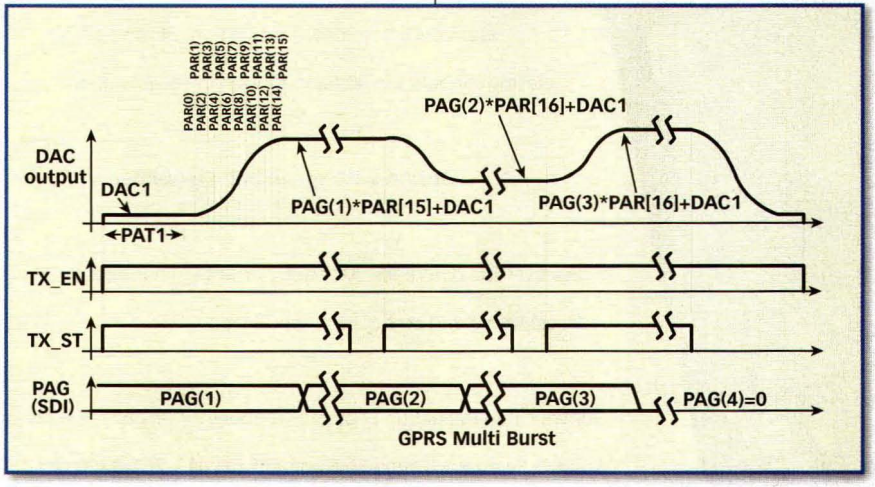
Following the analog polyphase filtering within the RF2708, frequency translation, and conversion to the digital realm, further filtering is performed by digital methods. The use of digital filters provides several benefits, including no group-delay variance; stable performance over voltage, temperature, and process tolerances; smaller chip area than analog filters; and lower power consumption when compared to analog filters. Digital filters also provide the ability to dynamically modify the channel bandwidth of the system to accommodate different modes of operation or varying baseband capabilities. The RF6001 includes eight selectable half-channel bandwidths from 75 to 135 kHz.

The RF6001's transmit section includes analog and digital I/Os (for use of the greatest number of baseband devices), a fractional-N synthesizer, a digital GMSK modulator, dual-power VCOs, and a 16-b PA ramp-control DAC. The RF6001 also includes integrated loop filters for the high- and low-band output VCOs in the transmit section and the receive VCO in the RF2708. The fractional-N synthesizer provides the core for the receive and transmit phased-locked loops (PLLs). It provides programmable reference frequency (R) and VCO (N) dividers. The N divider contains a dual-modulus prescaler and digitally spur-compensated fractional sequence generator for fine frequency steps (as small as 1.55 Hz for the digital GMSK modulator).

A deadzone/free-phase detector supports the fast switching times required for GPRS and EDGE. The phase detector aids low-jitter locking performance with the ability to hop frequencies and settle to within 100 Hz of a final frequency in only 125 μs. A lock-detect output pin is provided to validate proper PLL performance.

The digital GMSK modulator significantly simplifies the transmit section, eliminating the need for the translational loop circuitry seen in the majority of GSM/GPRS chip-set solutions. The modulator features low power consumption and low output noise. The complete transmit chain (excluding the PA) draws 45 mA compared to 70 to 90 mA for translational-loop systems and more for direct-launch RF modulators.

The VCOs achieve noise of better than −162 dBc/Hz offset 20 MHz from the carrier, eliminating the need for filters before or after the PA. This perfor-



2. The power-ramp capability of the RF6001 transceiver enables different output-power levels from slot to slot for GPRS applications.

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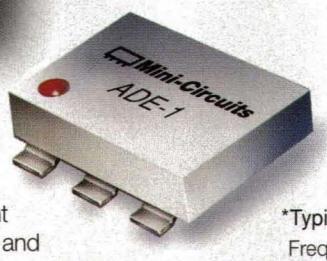

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Frequency LO/IF (MHz)	DC-500	5-1000
LO Level (dBm)	7	7
IP3 (dBm)	15	9
Conv. Loss (dB)	5.0	7.1
L-R Isolation (dB)	55	36
L-I Isolation (dB)	40	37
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mance enables direct connection from the VCO outputs to the PA inputs.

When using the transmit digital interface, digital non-return-to-zero (NRZ) data can be taken directly from the baseband device, eliminating the need for the baseband's modulator, Gaussian filter, and ADCs. Alternatively, the RF6001 has been designed to accept GMSK modulated analog I/Q signals from a baseband device.

The power-ramp DAC in the RF6001 reduces the control lines required between the baseband and the radio and lessens demands on the baseband processor. The DAC makes use of an efficient indirect closed-loop power-control scheme implemented in many of the company's PA modules (where the collector voltages for each stage can be regulated). The power-ramp waveform is based on 16-b settings in programmable registers within the RF6001. The waveforms for power ramp up and ramp down are symmetrical. The capability to change power levels in adjacent time slots supports GPRS operation. This enables the subsequent time-slot level to be set during the present slot enabling differing output power levels from slot to slot **(Fig. 2)**. The power-ramp DAC drives the V_{ramp} pin on the RF3133 PA, providing more than 35 dB of output-power control range. The RF3133 PA module offers maximum output power of +35 dBm for the EGSM band and +33 dBm for the DCS and PCS bands.

The RF6001 contains circuitry that can be configured with an external crystal to operate as a stand-alone 26-MHz reference oscillator, or as an input buffer for an externally generated 26-MHz clock. As an oscillator, the output can be configured to provide a 13- or 26-MHz clock output that can be used as the system clock.

A second 16-b DAC in the RF6001 can be used to provide control of a varactor diode, enabling the use of a lower-cost crystal (with less critical stability requirements).

To provide a truly "universal" solution, the Polaris chip set offers digital and analog input/output (I/O) interfaces in receive and transmit paths to support the broadest number of baseband devices. The digital interface provides three specific selectable modes of operation accommodating various baseband devices. Digital filtering in the receive path, and the digital GMSK modulator in the transmit path support digital interfaces between the baseband and the RF sections.

With digital interfaces, the ADCs and DACs in baseband devices can be eliminated, clearing the way for purely digital baseband devices. MRF

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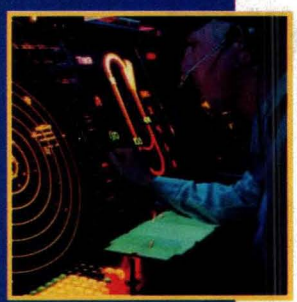

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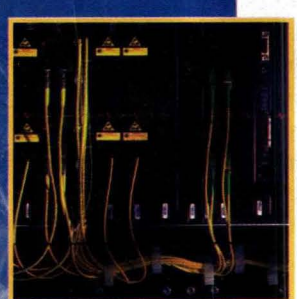

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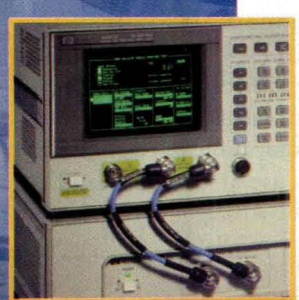

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Handset Amplifiers Arrive From New Source

A company long associated with workstations and personal computers (PCs) is now a supplier of high-efficiency PAs at cellular and PCS frequencies.

JACK BROWNE
Publisher/Editor

Power was not considered one of the strong points of silicon-germanium (SiGe) technology when the first devices were announced several years ago. But give credit to the technology's founder, IBM Microelectronics (Hopewell Junction, NY) for working to improve the process, and circuit design enhancements, enabling the development of the company's first line of handset power amplifiers (PAs) for cellular communications applications. The first three products can be used in Advanced Mobile Phone Service (AMPS), time-division-multiple-access (TDMA), and code-division-multiple-access (CDMA) designs operating from 824 to 849 MHz and from 1850 to 1910 MHz.

Cellular handset PAs were once thought to be the exclusive domain of gallium-arsenide (GaAs) technology, since early SiGe processes were limited in breakdown voltage. But IBM has succeeded in enhancing their SiGe materials and using a straightforward bipolar CMOS (BiCMOS) to fabricate the models IBM2017, IBM2018, and IBM2022 PAs for cellular and PCS frequencies and AMPS, TDMA, and CDMA use. The devices are fabricated in the company's Burlington, VT facility using the BiCMOS 5AM 0.5-μm SiGe process. They are designed for use with a single power supply; no reference voltage is needed.

These SiGe devices, which take advantage of the superior thermal conductivity of SiGe compared to GaAs, run cooler than GaAs devices for a particular output-power level, resulting in a measurable improvement in thermal reliability. Another advantage of SiGe technology is the capability of integrating analog and digital circuits on the same chip. As a result, the new amplifiers feature a fully integrated RF/bias/PA control network, as well as on-chip VSWR protection circuitry that allows each PA to handle load mismatches as severe as a 10.0:1 VSWR at +5 VDC collector voltage and full RF drive levels (input levels to +4 dBm). According to Dr. Teddy O'Connell, manager for wireless strategy and technical marketing, the new amplifiers offer "Si repeatability with GaAs performance." The amplifiers have been subjected to high-temperature RF stress testing to validate their long-term reliability and low junction temperatures.

The firm has actually announced three new amplifier products with the process, models IBM2022 for AMPS/TDMA IS-54 use at 800 MHz, IBM2018 for AMPS/CDMA IS-95 use at 800 MHz, and IBM2017 for personal-communications-services (PCS) CDMA I-95 use at 1900 MHz. The IBM2022 is a two-stage PA that can be

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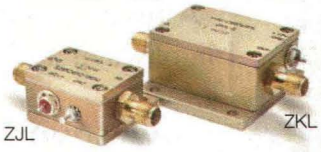
SPECIFICATIONS

Model	Freq (MHz)	Gain (typ) Midband (dB)	Gain (typ) Flat (±dB)	Max. P_{out}1 (dBm)	Dynamic Range (Typ @2GHz[2]) NF(dB)	Dynamic Range (Typ @2GHz[2]) IP3(dBm)	I(mA)[3]	Price $ea. (1-9)
ZJL-5G	20-5000	9.0	±0.55	15.0	8.5	32.0	80	129.95
ZJL-7G	20-7000	10.0	±1.0	8.0	5.0	24.0	50	99.95
ZJL-4G	20-4000	12.4	±0.25	13.5	5.5	30.5	75	129.95
ZJL-6G	20-6000	13.0	±1.6	9.0	4.5	24.0	50	114.95
ZJL-4HG	20-4000	17.0	±1.5	15.0	4.5	30.5	75	129.95
ZJL-3G	20-3000	19.0	±2.2	8.0	3.8	22.0	45	114.95
ZKL-2R7	10-2700	24.0	±0.7	13.0	5.0	30.0	120	149.95
ZKL-2R5	10-2500	30.0	±1.5	15.0	5.0	31.0	120	149.95
ZKL-2	10-2000	33.5	±1.0	15.0	4.0	31.0	120	149.95
ZKL-1R5	10-1500	40.0	±1.2	15.0	3.0	31.0	115	149.95

NOTES:
1. Typical at 1dB compression.
2. ZKL dynamic range specified at 1GHz.
3. All units at 12V DC.



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used for AMPS and TDMA applications from 824 to 849 MHz. It can operate with as little as 2-μA standby current, and delivers +31.5 dBm output power at 1-dB compression with 50-percent power-added efficiency (PAE) when biased for AMPS and +29.5-dBm output power with 40-percent PAE and 30.5-dB typical gain for TDMA operation. The gain varies only ±1 dB in AMPS applications from −30 to +85°C, and only ±0.8 dB for TDMA applications over the same temperature range. In TDMA designs, the adjacent-channel power ratio (ACPR) is specified at typically −29 dBc measured ±30 kHz from the carrier, while the alternate-channel power is typically −49 dBc when measured ±60 kHz from the carrier. This amplifier can also be operated in a low-power TDMA mode with +16-dBm output power, 28.5-dB gain, and 8-percent PAE, while maintaining the ACPR and alternate-channel-power performance of the higher-power TDMA mode. The IBM2022 is available in a 20-lead, 4-mm low-profile QFN package. The package has partial impedance matching for input and intermediate stages, with each requiring only one off-ship matching element. The output stage requires off-ship chip matching.

The IBM2018 is a two-stage amplifier module in a 6-mm-square LGA housing that can be used for IS-95 dual-mode (AMPS/CDMA) operation from 824 to 849 MHz. It delivers +31.5-dBm output power in AMPS systems and +28.5-dBm output power for CDMA use. The PAE is 50 percent in AMPS systems and 34 percent in CDMA systems. Supplied in a 6 × 6-mm LGA module, the amplifier features low standby current of less than 2 μA.

The IBM2017 is a three-stage device that is optimized for IS-95 PCS CDMA applications from 1850 to 1910 MHz. With on-chip log-controlled power shutdown bias control, the amplifier features +28.5- dBm output power and 32-percent PAE. The small-signal gain is typically 29 dB, with ±2-dB gain variation with temperature. The ACPR is at least −45 dBc measured ±1.25 MHz from the carrier, with alternate-channel power at levels of typically −51 dBc measured ±1.98 MHz from the carrier. The amplifier, which is supplied in a 16-lead LGA module, also offers a low-power CDMA mode with +12-dBm output power and typically 25-dB gain. The quiescent current is typically 85 mA in either operating mode. For more on IBM's SiGe process, or the availability of these amplifiers, visit the semiconductor portion of the company's website at www.chips.ibm.com.

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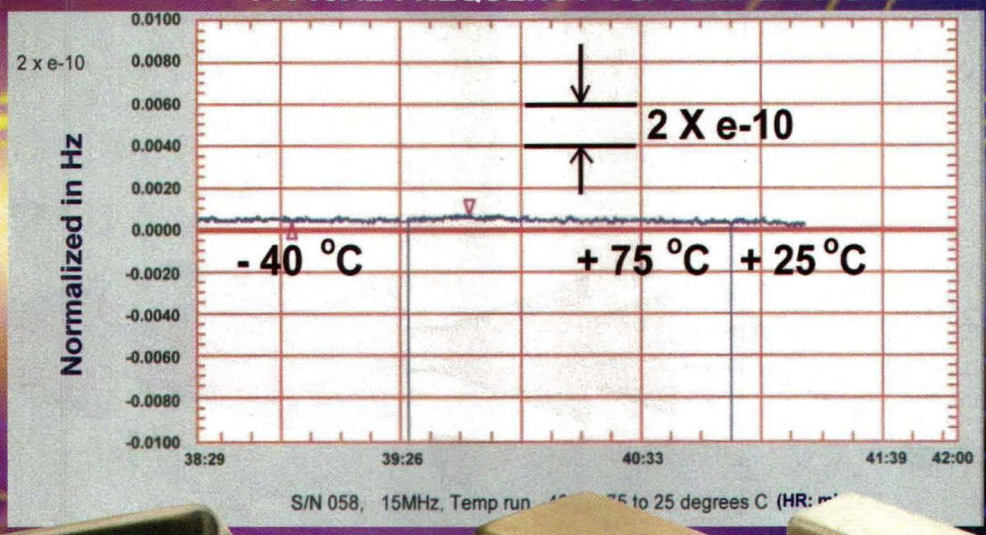

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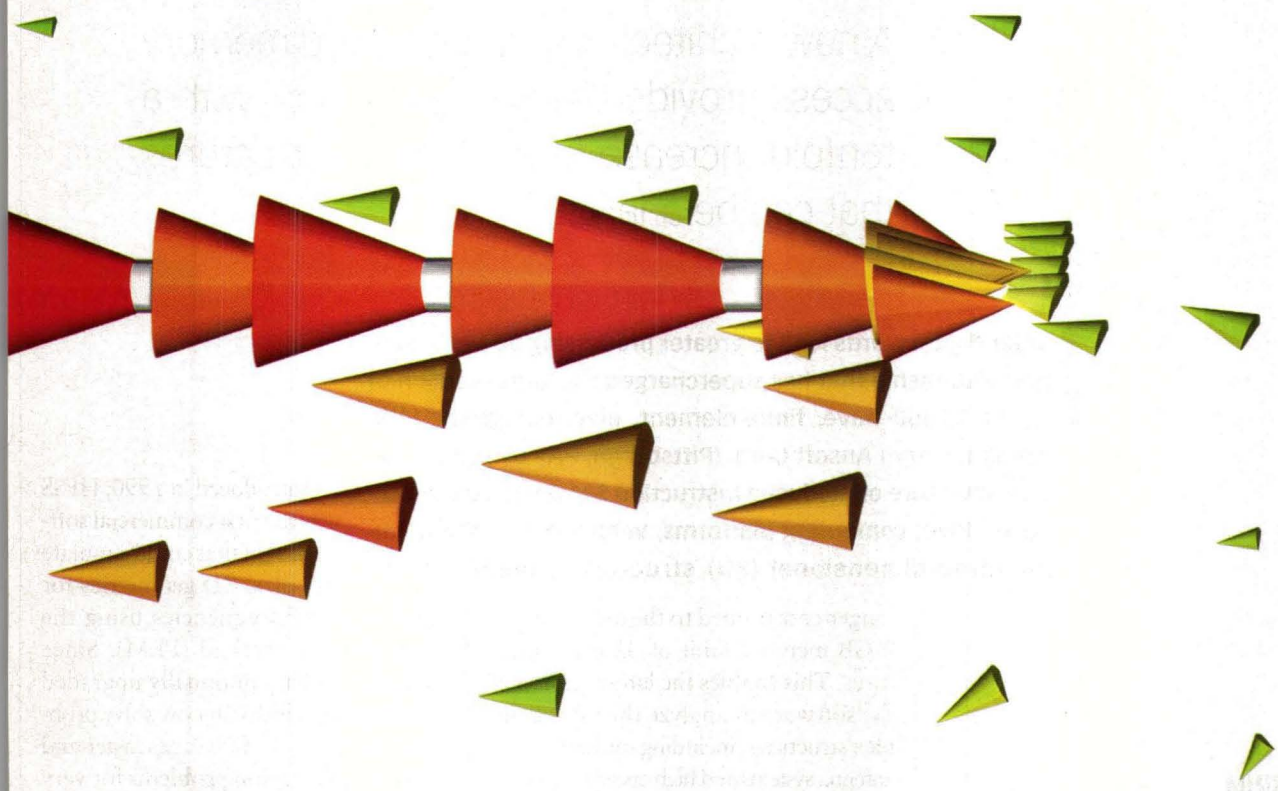

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EM Simulator Gets Boost With 64-b Code

A new architecture and increased memory access provide the HFSS simulator with a tenfold increase in the size of structures that can be analyzed.

Larger digital words lead to greater processing power, a simple relationship that has supercharged the latest version of the HFSS full-wave, finite-element, electromagnetic (EM) simulator from Ansoft Corp. (Pittsburgh, PA). By using a 64-b architecture on Reduced Instruction Set Computer (RISC)-based UNIX computing platforms, version 8.5 of the popular three-dimensional (3D) structure simulator is no longer constrained to the approximate 2-GB memory limit of 32-b architectures. This enables the latest version of the software to analyze the most complex structures, including multiple-array antenna systems and high-speed integrated-circuit (IC) packages.

BRAD BRIM
Product Marketing Manager for Microwave/RF Applications
Ansoft Corp., Four Station Square, Pittsburgh, PA 15219; (412) 261-3200, FAX: (412) 471-9427, Internet: www.ansoft.com.

HFSS enables engineers to design 3D, high-frequency structures, such as microwave filters, connectors, wave guides, IC packages, and antennas. HFSS 8.5 implements key sections of code using 64-b features, increasing its accessible memory space to accommodate the large memories typical of high-end workstations. The result is the industry's first commercial 64-b EM simulator, and a tenfold increase in the size and complexity of the structures HFSS can simulate. Ansoft will apply this same effort to its other design software in upcoming releases. The new version also can consider the frequency variation of any arbitrary material property and uses a significantly improved method for importing geometries from mechanical design packages, such as Pro/Engineer® and AutoCAD®.

Introduced in 1990, HFSS was the first commercial software tool that could simulate complex, 3D geometries for microwave/RF frequencies using the finite-element method (FEM). Since then, Ansoft has continually upgraded the software, which can now solve problems more than 15,000 times faster and address open-region problems for very complex shapes. In addition, HFSS can perform fast frequency sweeps for wideband simulation and can create Full-Wave Spice™ models to account for all frequency effects in time-domain SPICE circuit simulators. Optimetrics™, the software's optimization and sensitivity-analysis engine, allows designers to synthesize improvements by specifying design goals and investigating manufacturing changes to reduce costs. HFSS employs techniques such as automatic mesh generation, tangential-vector finite elements, and Adaptive Lanczos Pade Sweep (ALPS). Accuracy of simulation results is assured because the software automatically computes multiple adaptive solutions until a user-defined convergence criterion is met.

As designs have become more complex and frequencies have increased, a

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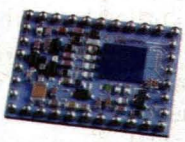

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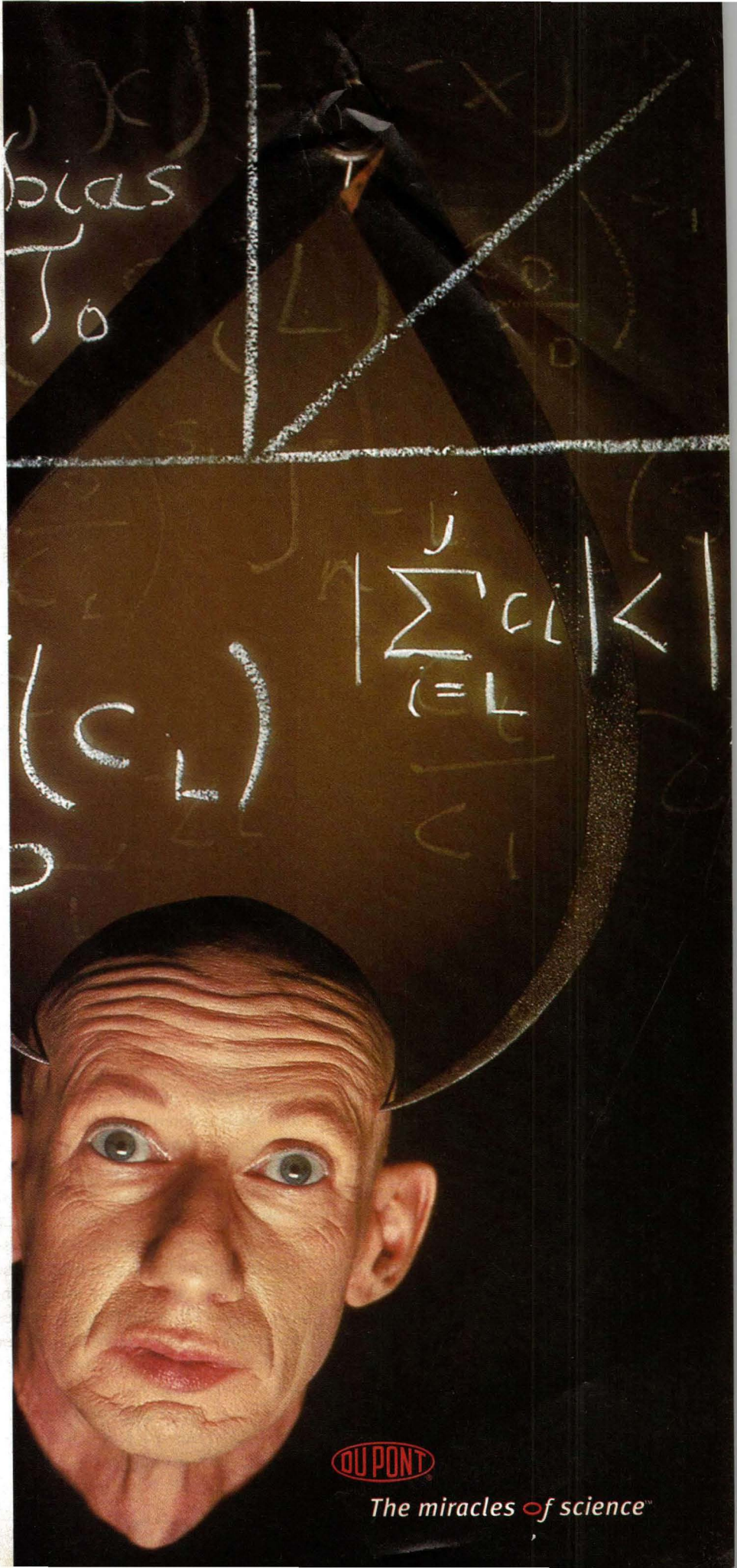

growing number of designers have found themselves constrained by the 2-GB limit of 32-b EM codes. For example, if a very complex antenna design required 1 GB of memory for highly accurate results, the use of 32-b code would preclude the possibility of analyzing this antenna as part of a four-element array. Similarly, for a 40-Gb/s OC-768 optical-communications system, designers could be limited by 32-b EM code to simulate only a small number of critical nets. With Version 8.5 of HFSS, more traces and irregularly shaped power/ground planes may be considered for more complete prediction of signal integrity for rise times of less than 25 ps.

There are basically two approaches Ansoft could have taken to enable a 64-b architecture in HFSS. The code could have simply been recompiled, applying 64-b features throughout. Although this would have been much easier and quicker for Ansoft, it would have significantly increased memory usage over its 32-b predecessor. Instead, Ansoft determined the key portions of HFSS that required modification to implement 64-b features. This effort rewards users with minimum impact on additional memory usage associated with 64-b codes.

Ansoft also enhanced HFSS 8.5 with the ability to automatically consider the broadband frequency variation of material properties, such as dielectric constant and loss. As HFSS computes the frequency-dependent response of a structure, it is now easy to evaluate arbitrary variations of any material property with respect to frequency. These spectral effects are then automatically considered in the time-domain response subsequently produced by Full-Wave Spice.

The consideration of frequency-dependent material properties can be critical at microwave/millimeter-wave frequencies. Notably, the characteristics of FR4, one of the most common printed-circuit-board (PCB) materials, are frequency-dependent at RF and microwave frequencies. The ability to consider the frequency-dependent material properties of FR4 for RF PCBs allows designers to reliably determine if their circuits will behave properly, or if more frequency-stable (though more expensive) microwave laminates are required to achieve the desired performance. These predictive design choices can save significant product manufacturing cost.

FR4 and other common PCB laminates are comprised of glass fibers and epoxy. The dielectric constant of the epoxy varies with frequency, as does the loss tangent of the glass. The impedance of printed lines depends directly on the dielectric constant of the surrounding layers of laminate. Consideration of this frequency-dependent effect may be important for increasingly popular controlled impedance designs, regardless of whether single-ended lines or differential pairs are used. Manufacturers of PCB laminates typically specify dielectric constant and loss tangent near DC and at a moderate microwave frequency. The properties most often vary linearly with respect to frequency in this range. For convenience of the user, these specified values and frequency behavior are the default properties applied by HFSS. For designs that include intentional frequency-dependent circuitry, such as tuning stubs or coupled lines used for constructing filters, the effects of material properties will assure first-pass success for that portion of the circuit.

The ability to consider very general variations of material properties with respect to frequency also enhances design capability for more esoteric applications, such as determining the behavior of plasmas or other materials for which the dielectric or loss properties vary significantly at microwave frequencies. HFSS supports this by allowing users to specify very general mathematical expressions for complex material properties.

Until now, simulation of these characteristics had to be performed one frequency point at a time. HFSS 8.5 requires the designer to only define the frequency-dependent material, and the software automatically performs the analysis over frequency, which can dramatically reduce engineering time.

Design of many microwave components begins in a mechanical engineering computer-aided-design (CAD) package or 3D solid modeling code. HFSS can import a solid model from these tools, effectively integrating mechanical and electrical design flows. HFSS directly imports ACIS (.sat) files common to many mechanical engineering CAD codes and other codes that use STEP or IGES formats to describe solid models. IGES and STEP files generated by mechanical engineering CAD codes will often have duplicate points, misaligned edges, or incorrect surface-normal vector definitions. These conditions make it impossible to create an adequate definition of geometry for EM simulation. HFSS 8.5 automates the ability to detect and fix these variances inherent in files created by mechanical design tools, a process known as "healing." A simple graphical-user interface (GUI) guides the user through this process. Should the conditions be beyond repair by the multiple healing techniques implemented in HFSS 8.5, a description of the solid objects and their condition is provided to support modification in the source mechanical-engineering CAD code.

HFSS 8.5 is the first version of the product to support a seamless transition to the Ansoft HFSS drawing environment from projects created with HFSS that are offered by Agilent Technologies (Santa Clara, CA). Ansoft licensed the Agilent version of HFSS software earlier this year with the goal of supporting Agilent's existing customers while gradually integrating them into Ansoft HFSS.

One of the most pressing questions to be asked after the alliance was formed was how to integrate projects created with Agilent HFSS into the Ansoft design environment. Ansoft HFSS can import Pro/Engineer and AutoCAD files from the most current versions of this software. HFSS 8.5 streamlines this process while retaining conventions familiar to users of Agilent HFSS. P&A: $41,900 and up; stock to 4 wks. **Ansoft Corp., Four Station Square, Pittsburgh, PA 15219; (412) 261-3200, FAX: (412) 471-9427, Internet: www.ansoft.com.**

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Size and weight must be minimized in space and aerospace applications. The tiniest gram of weight saved in these areas is pertinent to the product's success. When it comes to filters, StratEdge's (San Diego, CA) line of solid-state ceramic stripline filters replaces traditional microstrip filters that tend to weigh more and consume more space. Each filter is designed to customer specifications.

PETER STAVENICK
Managing Editor

Available as bandpass, lowpass, highpass, or combination designs ranging from 250 MHz to 18 GHz, the filters are very stable across the operating temperature range of −55 to +125°C. The filters' power versus frequency can be seen in **the figure.** They are constructed of 99.5 percent alumina or barium titanate, which has a high dielectric constant of 80. In the case of the stripline filters, a patented dimensionally stable manufacturing process was used.

The company has also used many unique processes when building its ceramic packages and filters. The key material is hardened or post-fired ceramic. Unlike other manufacturers, who use co-fired ceramic whose wider dimensional tolerances limit performance at high frequencies and have had only moderate success in the microwave frequency range, the post-fired hardened ceramic does not shrink.

Having manufactured filters for military and space applications along with phase-array antenna modules, the company offers good part-to-part repeatability while performing electrical and environmental testing for customers. The filters are suitable for Rxs and Txs and tuning is not required, which saves on cost. Since this series is self-contained, packaging is not needed.

Ranging from 200 to 300 at microwave frequencies, the typical designs range in size from 1.5 in. × 500 mils × 100 mils thick to as small as 310 × 100 × 30 mils thick. The filters have been designed with orders as high as 15 poles at 12 GHz, but mostly range from 4 to 15 pole.

Although the power-handling capabilities of these filters have not been fully explored, most designs should readily handle better than 5-W CW power and 100-W (short) pulsed power. **StratEdge, 4393 Viewridge Ave., San Diego, CA 92123; (858) 569-5000, FAX: (858) 560-6877, e-mail: info@strat edge.com, Internet: www.stratedge.com.**

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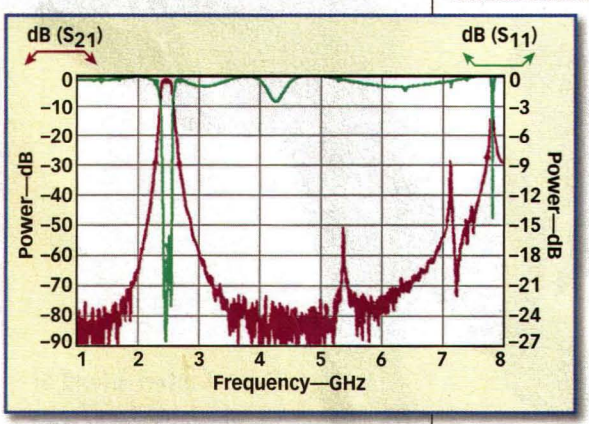
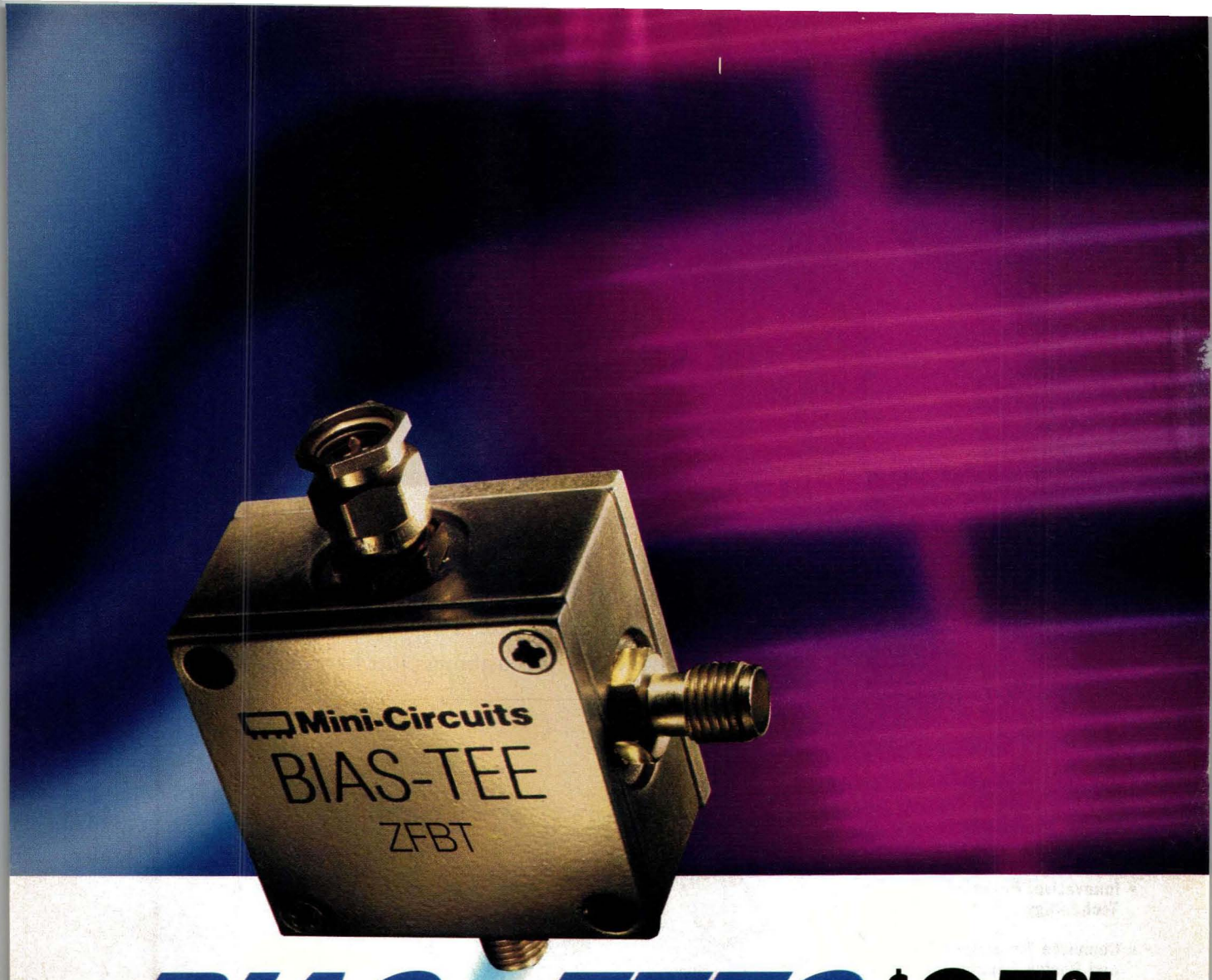


This filter graph shows power versus frequency.



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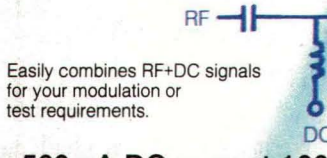

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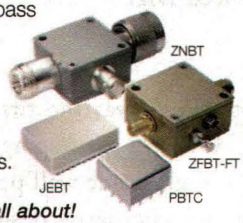

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▲ZFBT-4R2G	10-4200	0.15	0.6	0.6	32	40	50	1.13:1	59.95
▲ZFBT-6G	10-6000	0.15	0.6	1.0	32	40	30	1.13:1	79.95
▲ZFBT-4R2GW	0.1-4200	0.15	0.6	0.6	25	40	50	1.13:1	79.95
▲ZFBT-6GW	0.1-6000	0.15	0.6	1.0	25	40	30	1.13:1	89.95
▲ZFBT-4R2G-FT	10-4200	0.15	0.6	0.6	N/A	N/A	N/A	1.13:1	59.95
▲ZFBT-6G-FT	10-6000	0.15	0.6	1.0	N/A	N/A	N/A	1.13:1	79.95
▲ZFBT-4R2GW-FT	0.1-4200	0.15	0.6	0.6	N/A	N/A	N/A	1.13:1	79.95
▲ZFBT-6GW-FT	0.1-6000	0.15	0.6	1.0	N/A	N/A	N/A	1.13:1	89.95
★ZNBT-60-1W	2.5-6000	0.2	0.6	1.6	75	45	35	1.35:1	82.95
■PBTC-1G	10-1000	0.15	0.3	0.3	27	33	30	1.10:1	25.95
■PBTC-3G	10-3000	0.15	0.3	1.0	27	30	35	1.60:1	35.95
■PBTC-1GW	0.1-1000	0.15	0.3	0.3	25	33	30	1.10:1	35.95
■PBTC-3GW	0.1-3000	0.15	0.3	1.0	25	30	35	1.60:1	46.95
•JEBT-4R2G	10-4200	0.15	0.6	0.6	32	40	40	-	39.95
•JEBT-6G	10-6000	0.15	0.7	1.3	32	40	40	-	59.95
•JEBT-4R2GW	0.1-4200	0.15	0.6	0.6	25	40	40	-	59.95
•JEBT-6GW	0.1-6000	0.15	0.7	1.3	25	40	30	-	69.95

L = Low Range M = Mid Range U = Upper Range

NOTE: Isolation dB applies to DC to (RF) and DC to (RF+DC) ports.

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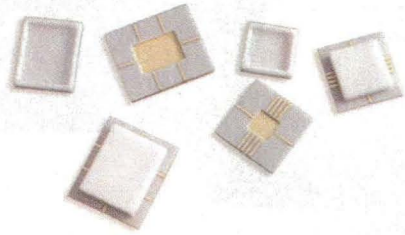

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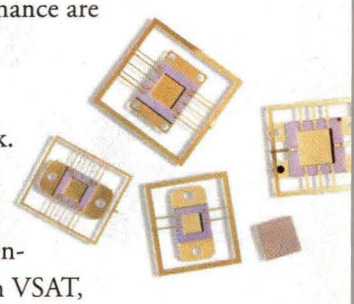

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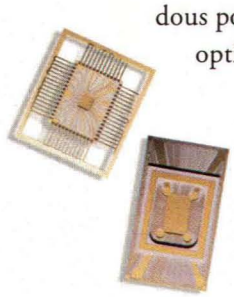

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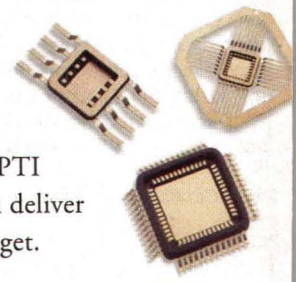

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MARKETING AND ADVERTISING STAFF

GROUP PUBLISHER
Craig Roth
(201) 393-6225
e-mail:croth@penton.com

SALES ASSISTANT
Judy Kollarik
(201)393-6218
e-mail: jkollarik@penton.com

DIRECT CONNECTION ADS
Joanne Reppas
(201) 666-6698
e-mail: jrepfran@aol.com

CLASSIFIED ADVERTISING
Loree Poirier
(216) 931-9201
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NEW YORK, NEW ENGLAND, SOUTHEAST, MIDWEST, MID-ATLANTIC, EASTERN CANADA
Paul Barkman
Regional Sales Manager
Penton Media, Inc.
611 Route #46 West
Hasbrouck Heights, NJ 07604
(908) 704-2460
FAX: (908) 704-2486
e-mail: pbarkman@penton.com

CALIFORNIA, NORTHWEST, SOUTHWEST, WESTERN CANADA
Gene Roberts
Regional Sales Manager
Penton Media, Inc.
6792 Almaden Road
San Jose, CA 95120
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Cell: (408) 640-7567
FAX: (408) 268-3415
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ITALY
Cesare Casiraghi
Via Napo Torriani 19/c
1-22100 Como - Italy
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GERMANY, AUSTRIA, SWITZERLAND
Friedrich K. Anacker
Managing Director
InterMedia Partners GmbH (IMP)
Deutscher Ring 40
42327 Wuppertal
Germany
Phone: 011-49-202-271-690
FAX: 011-49-202-271-6920

SPAIN
Luis Andrade, Miguel Esteban
Espana
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Sepulveda, 143-38
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Phone: 011-34-93-323-3031
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FRANCE
Emmanual Archambeaud
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EUROPEAN OPERATIONS
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Elan Marketing Group
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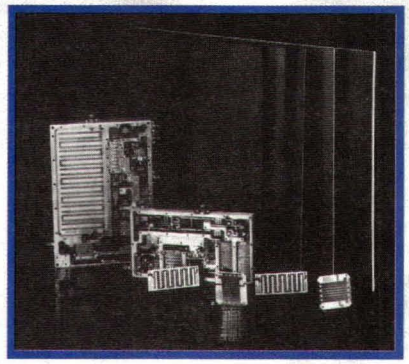

MORE THAN 14 YEARS AGO, a Special Report on microwave substrates compared and contrasted the benefits of hard ceramic and soft PTFE and Teflon base materials such as pure PTFE Cuflon from Polyflon Corp.

next month

Microwaves & RF April Editorial Preview

Issue Theme: Passive Components

News

This past February marked the 10th Anniversary of the original Wireless Symposium & Exhibition (now known as Wireless Systems 2002). Several hundred companies set up displays on the exhibit floor while thousands of attendees came to view new wireless products or further their education in the workshops and technical sessions. A special news report in April, co-authored by editors Peter Stavenick and Jack Browne, will present some of the technological and product highlights of the show, with a special focus on advances in passive components. In addition, show-floor interviews will reveal some of the thinking of customers and manufacturers for future prospects in wireless business.

Design Features

April opens with the insights of an engineer from Rockwell Collins on the design of a microstrip coupled-line bandpass filter. Also, an author from Analog Devices explores a multicarrier architecture for TD-SCDMA base-station transmitters. In addition, authors from Agilent Technologies and Anaren Microwave detail the design of a balanced LNA for cellular base stations, while Farron Dacus continues his series on the design of short-range, low-power radio systems, and an article series on LNAs from Maxim resumes with a chapter on device characterization.

Product Technology

The April Product Technology section will feature a line of broadband amplifiers with outstanding isolation properties from a leading supplier of high-frequency active and passive components. In addition, the Product Features section will examine a line of V-band components for wideband, license-free communications at 60 GHz, new series of two-, three-, and four-way power dividers/combiners based on a reliable multilayer circuit process, a low-cost spectrum-analyzer attachment for an oscilloscope, and a custom modular low-noise VCO.

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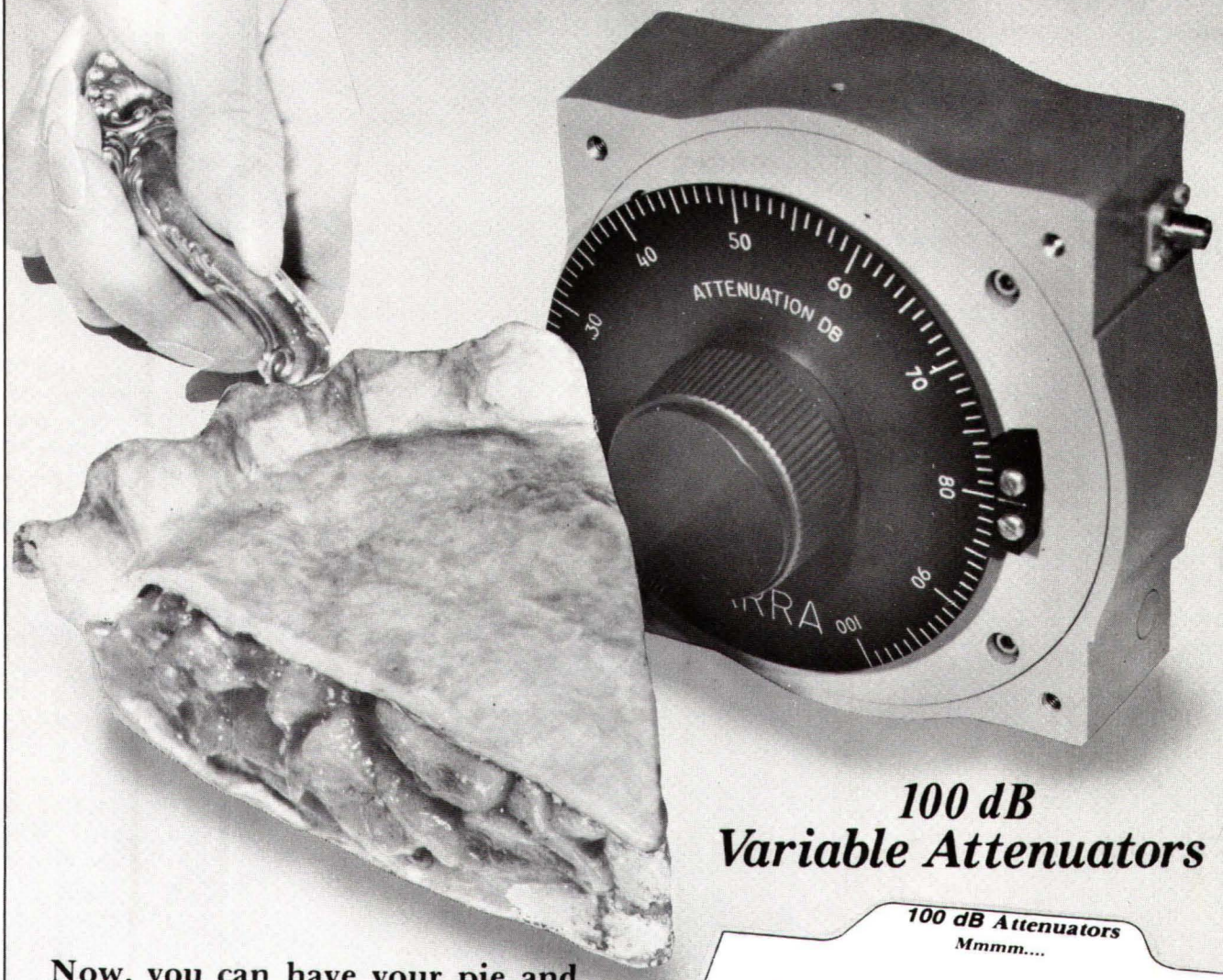

100 dB Variable Attenuators

100 dB Attenuators
Mmmm....

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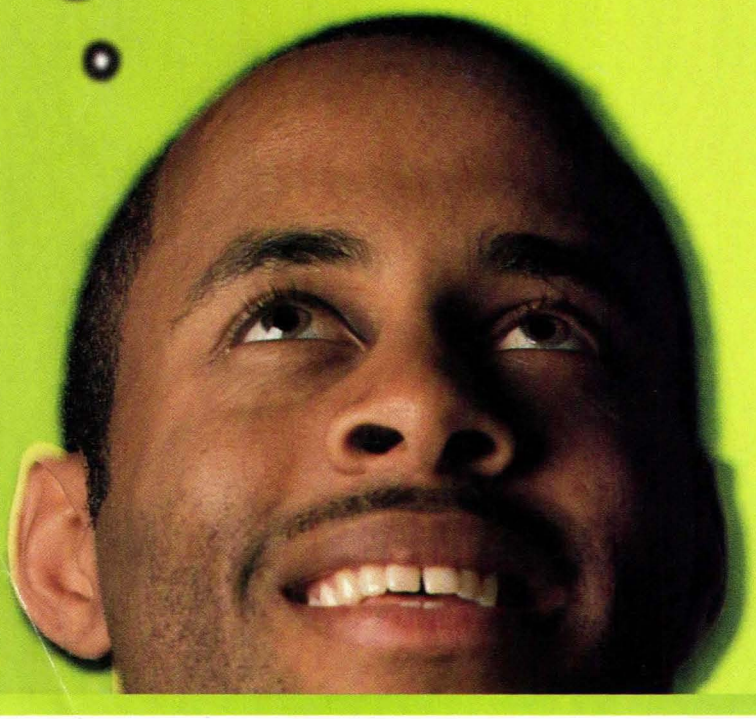

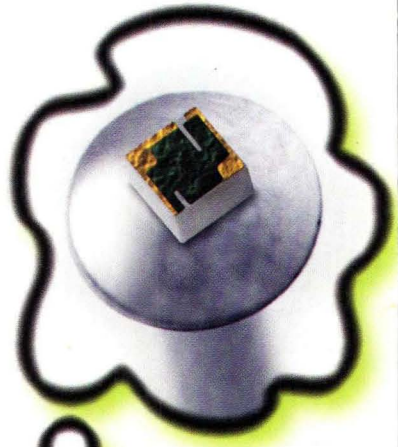

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